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NOTIFICATION OF ELECTION

(PCT Rule 61.2)

From the INTERNATIONAL BUREAU

To:

Commissioner
US Department of Commerce
United States Patent and Trademark
Office, PCT
2011 South Clark Place Room
CP2/5C24
Arlington, VA 22202
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in its capacity as elected Office

Date of mailing (day/month/year) 01 November 2000 (01.11.00)	
International application No. PCT/EP99/01618	**Applicant's or agent's file reference** 89350/VRD
International filing date (day/month/year) 10 March 1999 (10.03.99)	**Priority date** (day/month/year)
Applicant PIIRAINEN, Olli	

1. The designated Office is hereby notified of its election made:

   [X] in the demand filed with the International Preliminary Examining Authority on:

   08 September 2000 (08.09.00)

   [ ] in a notice effecting later election filed with the International Bureau on:

2. The election [X] was

   [ ] was not

   made before the expiration of 19 months from the priority date or, where Rule 32 applies, within the time limit under Rule 32.2(b).

The International Bureau of WIPO 34, chemin des Colombettes 1211 Geneva 20, Switzerland	Authorized officer Claudio Borton
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Form PCT/IB/331 (July 1992)

(54) Title: ESTIMATION OF DOPPLER SHIFT COMPENSATION IN A MOBILE COMMUNICATION SYSTEM

(57) Abstract

A system and method for estimating Doppler shift compensation in a mobile communication system is described. A channel impulse response determination circuit determines a channel impulse response for the channel on which the signal is received. A reference vector is generated using a set of known symbols and the channel impulse response, and an error vector is then generated by comparing the reference vector with received samples of the signal. In addition, the noise variance of the incoming signal can be determined. Then, a Doppler correction factor is estimated having a real part representing amplitude correction and an imaginary part representing phase correction.

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		(43) International Publication Date: 14 September 2000 (14.09.00)

(54) Title: ESTIMATION OF DOPPLER SHIFT COMPENSATION IN A MOBILE COMMUNICATION SYSTEM

(57) Abstract

A system and method for estimating Doppler shift compensation in a mobile communication system is described. A channel impulse response determination circuit determines a channel impulse response for the channel on which the signal is received. A reference vector is generated using a set of known symbols and the channel impulse response, and an error vector is then generated by comparing the reference vector with received samples of the signal. In addition, the noise variance of the incoming signal can be determined. Then, a Doppler correction factor is estimated having a real part representing amplitude correction and an imaginary part representing phase correction.

ESTIMATION OF DOPPLER SHIFT COMPENSATION
IN A MOBILE COMMUNICATION SYSTEM

The present invention relates to estimating a Doppler shift compensation in a mobile communication system.

In a mobile communication system, signals which are transmitted from mobile stations moving relative to a base station are subject to the well-known Doppler effect, which causes a frequency shift in the frequency received at the base station relative to that which was transmitted by the mobile station. This frequency shift is referred to herein as the Doppler shift. The Doppler shift is dependent upon the speed and direction of movement of the mobile station relative to the base station. Thus, the Doppler effect can provide an increase or a decrease in the frequency, depending on the direction of movement of the mobile station relative to the base station. The magnitude of the Doppler shift is dependent on the speed with which the mobile station is moving relative to the base station.

Another area of frequency offsets arises when the transmitter unit and the receiver unit are wrongly synchronised.

Existing mobile communication installations provide a form of Doppler compensation, in that the frequency detection circuitry within the base station which selects a particular signal on a particular channel can take into account a certain amount of Doppler shift in the signal.

AU 664626 relates to a method and circuit arrangement for compensating for the Doppler shift in a radio signal propagating between a base station and a mobile station when the mobile station approaches and moves past the base station. As a mobile station approaches the base station with a decreasing propagating time delay, the propagating time delay is integrated at intervals to determine how the propagating time delay varies with time. This information is used to effect a change in the radio signal

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frequency at a particular time to compensate for the sudden Doppler shift as the mobile station moves past the base station. In a TDMA system, the interrogation occurs at intervals equal to an integral number of time frames. In the GSM standard a time frame comprises eight consecutive time slots and a single transmission burst passes between a particular mobile station and base station in any one time frame.

Thus, in this method of Doppler compensation decisions are made in response to past and incoming signals to improve the reception at a future time, i.e. it is a reactive system. It would be desirable to implement a system which can actively compensate for Doppler shifts in an incoming signal in real time.

Moreover, due to fading a signal subject to the Doppler effect may have a variation in amplitude as well as frequency. It would be desirable to estimate the Doppler effect on amplitude as well as frequency.

According to one aspect of the invention there is provided a system for generating a Doppler correction factor for compensating for Doppler shift in a signal transmitted between a mobile station and a base station in a mobile communication system, the system comprising: a channel impulse response determination circuit for determining a channel impulse response for the channel on which the signal is received; reference circuitry for generating a reference vector from a set of known symbols and the channel impulse response; means for generating an error vector from the reference vector and received samples of the signal; means for determining noise variance of the incoming signal; and an estimator for generating a Doppler correction factor having a real part representing amplitude correction and an imaginary part representing phase correction, the estimator utilising the noise variance, the error vector, the samples of the received signal and the sampling times.

According to the GSM standard, a TDMA transmission burst

comprises a training sequence which is normally used to determine the channel impulse response for the channel on which the signal is received. This is done by a convolution of the received training sequence with a stored version of the training sequence. Thus, the training sequence can constitute the set of known symbols for use in generating the reference vector. Alternatively, the set of known symbols can comprise estimated symbols from the received signal samples.

In a TDMA system, the signal comprises a sequence of transmission bursts in respective time slots. The Doppler correction factor can be generated for each transmission burst and used to correct signal samples in that burst.

In addition for correcting for the Doppler effect, the Doppler correction factor also corrects for frequency offsets arising from wrong frequency synchronisation between transmitter and receiver units.

According to a further aspect of the present invention there is provided a method for compensating said Doppler shift in a signal transmitted between a mobile station and a base station in a mobile communication system, the method comprising: determining a channel impulse response for the channel on which the signal is received; generating a reference vector from a set of known symbols in the channel impulse response; generating an error vector from the reference vector and received samples of the signals; determining noise variance of the incoming signals; estimating a Doppler correction factor having a real part representing amplitude correction and an imaginary part representing phase correction, the estimating step using the noise variance, the error vector, the samples of the received signals and the sampling times; and applying the Doppler correction factor to the received signal samples thereby to compensate for Doppler shift.

For a better understanding of the present invention and to show

how the same may be carried into effect, reference will now be made by way of example to the accompanying drawings in which:

Figure 1 is a diagram of a signal burst in a mobile communication system;

Figure 2 is a block diagram of circuitry for implementing Doppler shift compensation; and

Figure 3 is a diagram illustrating the Estim β block.

Figure 1 illustrates a normal burst in a mobile communication system according to the GSM standard. This figure represents a burst received at a base station. For a TDMA system according to the GSM standard, mobile stations transmit bursts as modulated signals on frequency channels allocated by a base station controller. One frequency channel may support up to eight bursts, each burst associated with a respective call, where each call is allocated a time slot in which to send the burst. Further details of a TDMA system according to the GSM standard are not described herein because they are known to a person skilled in the art.

The normal burst contains two packets of 58 bits (DATA) surrounding a training sequence (TRS) of 26 bits. Three tail bits (TS) are added at each end of the normal burst. The training sequence (TRS) is a predetermined sequence of bits which is sent by the mobile station (MS) and is known at the base station controller (BSC). In Figure 1, it is assumed that a received signal sample y_k represents the k_{th} bit of the burst, in this instance the start of the training sequence. The training sequence is utilised at the base station controller to estimate the impulse response of the channel over which the burst is sent, and, at least in one embodiment of the invention to estimate a Doppler compensation factor. The actual information which is transmitted is located in the data bits (DATA) of the burst.

As explained earlier, the environment through which a signal passes from a mobile station to a base station can vary

considerably, depending, amongst other things, on the distance
between the mobile station and the base station, and interference
caused by buildings and other structures in the area.   As a
result, the signal strength and signal quality of the signal
received at the base station varies widely. Moreover, for moving
mobile stations, the signal received by the base station is
subject to a Doppler shift which should be corrected.

Figure 2 illustrates a circuit 1 suitable for implementing a
Doppler compensation in a GSM system.   It should be understood
that the various blocks in Figure 2, although illustrated as
separate interconnected entities, do not necessarily represent
separate physical entities, but are intended to represent
diagrammatically the various steps which are carried out. The
blocks could be implemented as circuits or a suitably programmed
microprocessor may effect each of the functions which is
individually assigned to the blocks.

An antenna 20 receives signals 11 from the mobile stations MS.
The antenna 20 is connected to RF circuitry 22. This circuitry
22 operates on the received burst to downshift the frequency to
the baseband frequency and to sample the burst to provide from
the analogue signal digital sampled values y_i where i is the
sample index.   The RF circuitry 22 outputs the sampled values
$y_0 \ldots y_n$ for each burst together with the sampling times $t_0 \ldots t_n$,
typically at the expected bit rate of the transmitted signal.
The output of RF circuitry 22 is supplied along line 24 to a
channel impulse response (C.I.R.) block 10, to a variance and
error calculator 16 to filtering and equalisation circuitry 12,
to an Estim β block 36 for estimating a Doppler correction factor
and to transforming circuitry 40 to enable the application of the
Doppler correction factor to the burst.

A memory 32 holds the training sequence x which is the
predetermined sequence of bits which is sent by the mobile
station MS as a training sequence and received at the base
station and converted into samples $y_{TRS} = Y_k \rightarrow y_{k+25}$. The reference

training sequence x is supplied to a reference generator 14 and
to the channel impulse response (C.I.R.) block 10. The reference
generator 14 also receives the estimated channel impulse response
h from the channel impulse response block 10.

The C.I.R. block 10 receives the burst, including the received
training sequence y_{TRS} and calculates an estimated channel impulse
response h for example by calculating the cross correlation
between the received training sequence y_{TRS} and the known training
sequence x according to equation 1. There are other types of
estimation methods which could be used.

$$h = \text{xcorr } (x, y_{TRS}) \qquad\qquad \text{(equation 1)}$$

It will be appreciated that, prior to effecting the cross
correlation, the known training sequence x, which is stored in
digital form, is i,q modulated similarly to the manner in which
the training sequence has been modulated at the MS for
transmission, according to the GSM standard. The cross
correlation is done in a known manner to produce a channel
impulse response in the form of five tap values $(h(j)_{j=0to4})$.

As is known, the estimated impulse response h is used to
calculate the expected estimate of the data in the received
burst, as though the data has been subject to the same average
noise.

The C.I.R. block also generates timing advance information τ,
which is used to determine where in the allocated time slot the
received burst is located.

For each burst, the estimated channel impulse response h for that
burst is calculated by the CIR block 10 and is supplied to
filtering/equalising circuitry which allows the data, DATA(r),
in that burst to be recovered. As is known, the
filtering/equalising circuit 12 receives the channel impulse
response h and timing information τ for the received burst to

allow the signal to be demodulated, filtered and decoded, to recover the data in a known manner.

The reference generator 14 produces a reference vector, z, which is calculated using the convolution of the impulse response and the known training sequence. The reference generator 14 performs the following calculation:

$$z = h*x \qquad\qquad \text{(equation 2)}$$

where x represents the known symbol values depending o the used modulation.

z_k represents the kth sample of the vector z and z denotes a diagonal matrix of the vector samples z_k, that is, z=xh, where x is the diagonal matrix of the known training sequence symbols.

The reference vector z is supplied from the reference generator to the variance and error calculator 16. As described above, the variance calculator also receives the samples y_i, including the received training sequence y_{TRS}. The variance calculator calculates a variance var (σ^2) according to the following equation:

$$var = \frac{\sum_{i=4}^{25}(|y_i - z_i|^2))}{ref\_length} \qquad\qquad \text{(equation 4)}$$

The term ref_length is a constant representing the length of the reference vector, z. This is calculated by multiplying the number of samples (22) by the bit separation.

In equation 4, the values of y_i are the sampled values of the received training sequence for the burst $(y_k \rightarrow y_{k+25})$.

The variance and error calculator also generates an error vector

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z-y (i.e. error values z_i-y_i for each sample).

The error vector z-y and the variance σ^2 are supplied to the Estim β block 36. The Estim β block 36 also receives the signal samples and times y_i, t_i from the RF circuitry 22. The Estim β block can be implemented as a processor executing a program or by other suitable means. It operates to generate a Doppler compensation factor β which is a complex number having real and imaginary parts. Thus, β may be represented as:

$$\alpha + j\phi t$$

where α represents amplitude variations due to fading and ϕ is the Doppler angle. The Estim β block 36 also includes a memory for holding values required to generate the Doppler correction factor β, including C_β, $E(\beta)$ and $E(z-y)$. These terms are explained later with reference to Figure 3.

The Doppler correction factor β together with the signal samples y_i and times t_i for each burst are supplied to the transforming circuitry 40 to implement Doppler correction according to the following equation:

$$Y_{n-compensated} = Y_n(1 - \beta t_n).$$

Another implementation of the transforming circuitry 40 is to separate the amplitude and phase components into separate correction steps. The phase can be corrected by multiplying the samples with $e^{-j*imag(\beta)t_n}$ and the amplitude can be corrected in a separate step:

$$Y_{n-phase-compensated} = Y_n * e^{-j*imag(\beta)t_n}$$
$$Y_{n-compensated} = Y_{-phase-compensated}(1 - real(\beta)t_n) \qquad \text{(equation 5)}$$

According to the above, $imag(\beta) = \phi$, and $real(\beta) = \alpha$. Thus, the received signal is either amplified or alternated during the burst depending whether $real(\beta)$ is negative or positive.

This generates a set of compensated signal samples y_{comp} which can be fed to the filtering/equalising circuitry 12 to generate refined values for the data of the burst.

As an alternative to the above, in which the training sequence x is used to supply known symbols as a reference, bits estimated in an equalisation circuit 30 can be used. The equalisation circuit 30, for example of Viterbi equaliser, receives the filtered, demodulated and equalised signal DATA (r) from the filtering and equalisation circuitry 12. The equalisation circuit 30 operates on a part of the data sequence data of the burst (that part having been derived from Estim.block in Figure 1) to estimate output bits which were sent from the mobile station. This output is referred to herein as estim_bits. The equalisation circuit 30 operates to make decisions of the bits as in known mobile communication systems and thus it will not be described further herein. These estimated bits can be used in place of the training sequence to supply known values for calculating the reference vector z. The estim_bits are supplied to a reference circuit 34 with the channel impulse response h which generates an error vector z'-y, where z'=estim_bits*h.

The theory underlying the generation of the Doppler correction factor β will now be outlined.

The problem of (pure) Doppler frequency estimation can be written as:

Xh=Yr+w, where r is a rotation vector,
Y is a diagonal sample matrix, and
w is noise.

$$Y=\begin{pmatrix} y_0 & 0 & \cdots & 0 \\ 0 & y_1 & \cdots & 0 \\ \cdots & \cdots & \cdots & \cdots \\ 0 & \cdots & \cdots & y_{n-1} \end{pmatrix}, r=\begin{pmatrix} e^{j\phi t}{}_0 \\ e^{j\phi t}{}_1 \\ \\ e^{j\phi t}{}_{n-1} \end{pmatrix}$$

Xh corresponds to convolution between known symbols (diagonal matrix x of training sequence symbols) and impulse response h.

So, in the Doppler frequency estimation, the Doppler angle ϕ is to be estimated. This problem is nonlinear and therefore a bit tricky to be solved, particularly for implementation in real time. A linearisation step can be performed assuming $e^{j\phi t_0} \approx 1 + j\phi t_0$. Due to fading the required Doppler frequency correction cannot be assumed to be constant in amplitude.

If the term $j\phi t_0$ has also a real component, the amplitude variations can roughly estimated. Thus, the term $j\phi t_0$ is replaced by a more general assumption $1 + \beta t_0$, where β is complex $(\alpha + j\phi t)$.

The linear problem can be written as (for simplification $z = Xh = [z_0 z_1 \ldots z_{n-1}]^T$).

$$z_i = y_i (1 + \beta t_i) + w_i,$$

which can be reformatted in matrix notation as

$$z - y = p\beta + w \qquad\qquad\qquad \text{(equation 6)}$$

where $p = [y_0 t_0 y_1 t_1 \ldots y_{n-1} t_{n-1}]^T$ and β is to be estimated.

A solution to equation 6 can be found by using an LMMSE estimator

$$\beta = E(\beta) + C_{\beta(z-y)} C_{(z-y)(z-y)}^{-1} (z - y - E(z-y))$$

where $E(\beta)$ is an expected value for β, $C_{\beta(z-y)}$, $C_{(z-y)(z-y)}$ are covariance matrixes between the two elements, viz $E(\beta(z-y)^H)$ and $E((z-y)(z-y)^H$ and $E(z-y)$ is the expected error in the received burst.

In this phase it can be assumed that the noise is white gaussian noise having a variance of δ^2 and after some manipulation

$$\beta = E(\beta) + p^H(z-y-E(z-y))/(\delta^2/C_\beta + p^Hp), \text{ where} \quad \text{(equation 7)}$$
$$C_\beta = E((\beta-E(\beta))(\beta-E(\beta))^H) \quad \text{(equation 8)}$$

p^H is the complex conjugate transpose matrix of p.

Equation 7 is used in the Estim β block 36 to generate β. The Estim β block is illustrated schematically in Figure 3. In practice, it can be implemented as software executed by a processor with a memory for holding the required data. An Estimate_β block 50 operates on the data (Y_i, t_i) received in each time slots to generate a value for β which is used by the transforming circuitry 40. An Acc β block 52 accumulates and averages β values over preceding time slots to generate $E(\beta)$. The Acc B function can be implemented by an adaptive filter. An Acc $(z-y)$ block 54 generates an expected error $E(z-y)$. To do this, it can use values for z-y from preceding time slots, or $E(\beta)$ from the Acc β block 52 to estimate a frequency error. A Gen C_β block 56 produces a value for C_β. A number of possibilities exist for the parameters of equation 7.

If $E(\beta)$ is set to zero, and δ^2/C_β is also set to zero, equation 7 resolves to a normal linear regression.

As mentioned above, it is intended that β will be generated for each time slot. The expected value $E(\beta)$ can be generated by calculating the average of estimated β values from previous time slots. Thus, these values can be retained as they are calculated for each time slot, and used to produce an average value held in the $E(\beta)$ block in the Estim β block 36.

Another simplification is to assume that the real part of $E(\beta)$ is zero. This is a reasonable assumption because, in typical fading channel situations, no long time amplitude is expected. Thus, it is adequate to perform long time averages only in respect of error components of frequency. Also, the real part could be assumed to be zero if only phase variations are to be corrected.

It is possible to avoid having to calculate the expected error $E(z-y)$ by derotating the received samples y_i by $e^{-j*imag(\beta)t}$. For this purpose, β can be an earlier estimated value of β from a previous time slot. In this case, equation 7 changes to

$$\beta = E(\beta) + \mathbf{p}^H(\mathbf{z}-\mathbf{y}) / (\delta^2/C_\beta + \mathbf{p}^H\mathbf{p}) \qquad \text{(equation 9)}$$

Thus, this requires a preprocessing step (see 50a in Figure 3) in which the sample values y_i are derotated by $e^{-j*imag(\beta)t}$ prior to forming the vector $\mathbf{p}$.

The value for $C(\beta)$ can be generated by the GEN C_β block using different approaches. It can be a constant value. I can be estimated using the information of previous time slots, using for example equation 8. It can be made a function of the expected value of β and/or the noise variance δ^2. It can be estimated using an adaptive filter, or any combination of the preceding.

The technique is particularly suitable for short delay spread channels and pure frequency errors.

CLAIMS:

1.   A system for generating a Doppler correction factor for
compensating for Doppler shift in a signal transmitted between
a mobile station and a base station in a mobile communication
system, the system comprising:

      a channel impulse response determination circuit for
determining a channel impulse response for the channel on which
the signal is received;

      reference circuitry for generating a reference vector from
a set of known symbols and the channel impulse response;

      means for generating an error vector from the reference
vector and received samples of the signal;

      means for determining noise variance of the incoming signal;
and

      an estimator for generating a Doppler correction factor
having a real part representing amplitude correction and an
imaginary part representing phase correction, the estimator
utilising the noise variance, the error vector, the samples of
the received signal and the sampling times.

2.   A system according to claim 1, wherein the Doppler
correction factor β is generated according to the following
equation:

$$\beta = E(\beta) + \mathbf{p}^H(\mathbf{z} - \mathbf{y} - E(\mathbf{z} - \mathbf{y})) / (\delta^2/C_\beta + \mathbf{p}^H\mathbf{p})$$

wherein $E(\beta)$ is an expected value of β,

$\mathbf{p}^H$ is a complex conjugate transpose matrix of $\mathbf{p}$, where $\mathbf{p}$ is the
vector $(y_i, t_i)^T$,

$\mathbf{z} - \mathbf{y}$ is the error vector,

$E(\mathbf{z} - \mathbf{y})$ is the expected error in the received samples,

δ^2 is the noise variance, and

C_β is a constant.

3.   A system according to claim 2, wherein $E(\beta)$ and δ^2/C_β are set to zero.

4.   A system according to claim 1, 2 or 3, which comprises means for applying a Doppler correction to the received signal samples using the Doppler correction factor β.

5.   A system according to any preceding claim wherein the signal comprises a transmission burst in a time slot in a TDMA mobile communication system, and wherein a new Doppler correction factor β is generated for each time slot.

6.   A system according to claim 5, which comprises means for estimating $E(\beta)$ as an average of generated values for β from previous time slots.

7.   A system according to claim 5, wherein the real part of $E(\beta)$ is set to be zero.

8.   A system according to any preceding claim, which comprises means for derotating the received signal samples by $e^{-j*imag(E(\beta))t}$, where $E(\beta)$ is an expected value for β the derotated received signal samples then being used to generate the Doppler correction factor β in the present time slot according to the following equation:

$$\beta = E(\beta) + \mathbf{p}^H(\mathbf{z}-\mathbf{y})/(\delta^2/C_\beta + \mathbf{p}^H\mathbf{p}).$$

9.   A method for compensating said Doppler shift in a signal transmitted between a mobile station and a base station in a mobile communication system, the method comprising:
        determining a channel impulse response for the channel on which the signal is received;
        generating a reference vector from a set of known symbols in the channel impulse response;

generating an error vector from the reference vector and received samples of the signals;

determining noise variance of the incoming signals;

estimating a Doppler correction factor having a real part representing amplitude correction and an imaginary part representing phase correction, the estimating step using the noise variance, the error vector, the samples of the received signals and the sampling times; and

applying the Doppler correction factor to the received signal samples thereby to compensate for Doppler shift.

10.   A method according to claim 9, wherein the signal comprises a sequence of transmission bursts in a TDMA mobile communication system, and the Doppler correction factor is generated for each transmission burst.

11.   A method according to claim 9 or 10, wherein the Doppler correction factor β is generated according to the following equation:

$$\beta = E(\beta) + \mathbf{p}^H (\mathbf{z}-\mathbf{y}-E(\mathbf{z}-\mathbf{y})) / (\delta^2/C_\beta + \mathbf{p}^H\mathbf{p})$$

wherein $E(\beta)$ is an expected value of β,

$\mathbf{p}^H$ is a complex conjugate transpose matrix of $\mathbf{p}$, where $\mathbf{p}$ is the vector $(y_i, t_i)^T$,

$\mathbf{z}-\mathbf{y}$ is the error vector,

$E(\mathbf{z}-\mathbf{y})$ is the expected error in the received samples,

δ^2 is the noise variance, and

C_β is a constant.
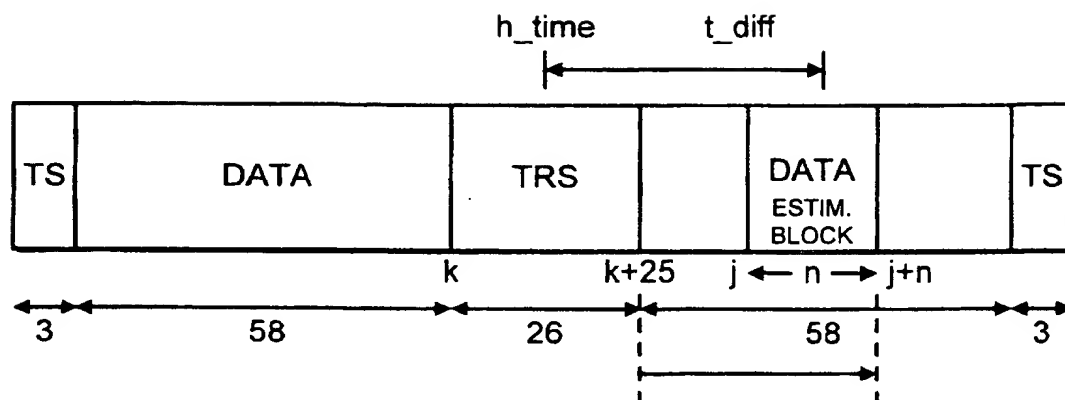
FIG. 1

FIG. 3

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FIG. 2

PCT
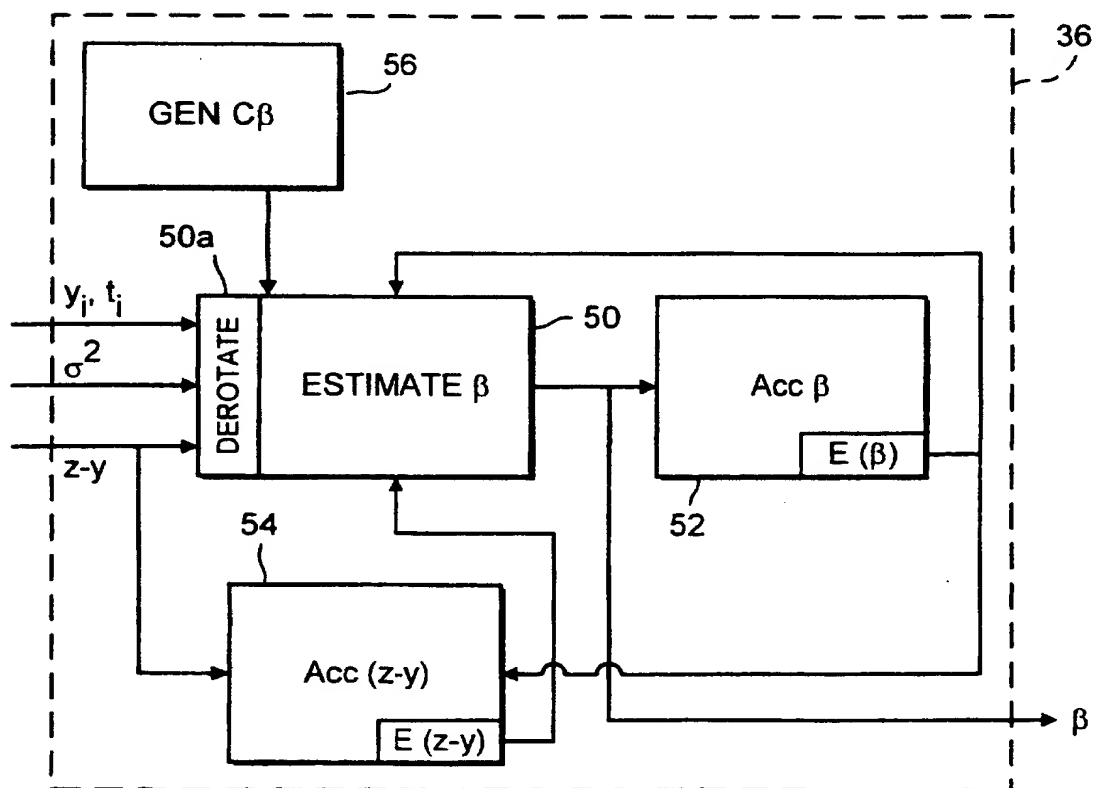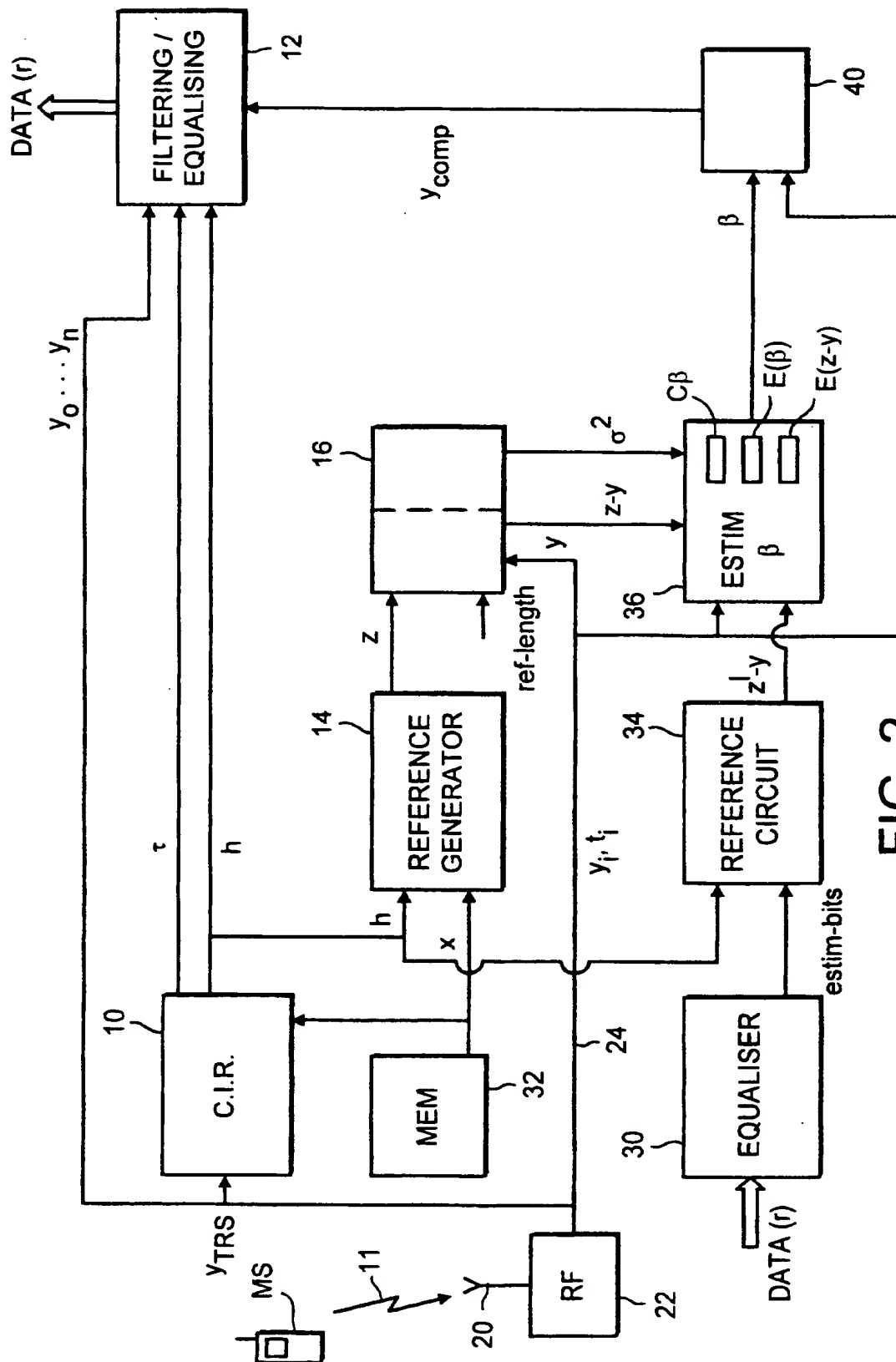
INTERNATIONAL PRELIMINARY EXAMINATION REPORT

(PCT Article 36 and Rule 70)

14

Applicant's or agent's file reference 89350/VRD	**FOR FURTHER ACTION**	See Notification of Transmittal of International Preliminary Examination Report (Form PCT/IPEA/416)
International application No. PCT/EP99/01618	International filing date *(day/month/year)* 10/03/1999	Priority date *(day/month/year)* 10/03/1999

International Patent Classification (IPC) or national classification and IPC H04B7/01

Applicant NOKIA NETWORKS OY et.al.

1. This international preliminary examination report has been prepared by this International Preliminary Examining Authority and is transmitted to the applicant according to Article 36.

2. This REPORT consists of a total of 5 sheets, including this cover sheet.

    ☐ This report is also accompanied by ANNEXES, i.e. sheets of the description, claims and/or drawings which have been amended and are the basis for this report and/or sheets containing rectifications made before this Authority (see Rule 70.16 and Section 607 of the Administrative Instructions under the PCT).

    These annexes consist of a total of sheets.

3. This report contains indications relating to the following items:

    I   ☒ Basis of the report
    II   ☐ Priority
    III ☐ Non-establishment of opinion with regard to novelty, inventive step and industrial applicability
    IV   ☐ Lack of unity of invention
    V   ☒ Reasoned statement under Article 35(2) with regard to novelty, inventive step or industrial applicability; citations and explanations suporting such statement
    VI   ☐ Certain documents cited
    VII ☒ Certain defects in the international application
    VIII ☐ Certain observations on the international application

Date of submission of the demand 08/09/2000	Date of completion of this report 18.06.2001
Name and mailing address of the international preliminary examining authority: European Patent Office D-80298 Munich Tel. +49 89 2399 - 0 Tx: 523656 epmu d Fax: +49 89 2399 - 4465	Authorized officer Burghardt, G Telephone No. +49 89 2399 8979

I. Basis of the report

1. With regard to the **elements** of the international application *(Replacement sheets which have been furnished to the receiving Office in response to an invitation under Article 14 are referred to in this report as "originally filed" and are not annexed to this report since they do not contain amendments (Rules 70.16 and 70.17)):*
   Description, pages:

   1-12   as originally filed

   Claims, No.:

   1-11   as originally filed

   Drawings, sheets:

   1/3-3/3   as originally filed

2. With regard to the **language,** all the elements marked above were available or furnished to this Authority in the language in which the international application was filed, unless otherwise indicated under this item.

   These elements were available or furnished to this Authority in the following language: , which is:

   ☐   the language of a translation furnished for the purposes of the international search (under Rule 23.1(b)).

   ☐   the language of publication of the international application (under Rule 48.3(b)).

   ☐   the language of a translation furnished for the purposes of international preliminary examination (under Rule 55.2 and/or 55.3).

3. With regard to any **nucleotide and/or amino acid sequence** disclosed in the international application, the international preliminary examination was carried out on the basis of the sequence listing:

   ☐   contained in the international application in written form.

   ☐   filed together with the international application in computer readable form.

   ☐   furnished subsequently to this Authority in written form.

   ☐   furnished subsequently to this Authority in computer readable form.

   ☐   The statement that the subsequently furnished written sequence listing does not go beyond the disclosure in the international application as filed has been furnished.

   ☐   The statement that the information recorded in computer readable form is identical to the written sequence listing has been furnished.

4. The amendments have resulted in the cancellation of:

   ☐   the description,   pages:

   ☐   the claims,   Nos.:

☐ the drawings,   sheets:

5. ☐ This report has been established as if (some of) the amendments had not been made, since they have been considered to go beyond the disclosure as filed (Rule 70.2(c)):

   (Any replacement sheet containing such amendments must be referred to under item 1 and annexed to this report.)

6. Additional observations, if necessary:

V. Reasoned statement under Article 35(2) with regard to novelty, inventive step or industrial applicability; citations and explanations supporting such statement

1. Statement

Novelty (N)	Yes:	Claims	1-11
	No:	Claims	
Inventive step (IS)	Yes:	Claims	1-11
	No:	Claims	
Industrial applicability (IA)	Yes:	Claims	1-11
	No:	Claims	

2. Citations and explanations
   see separate sheet

VII. Certain defects in the international application

The following defects in the form or contents of the international application have been noted:
see separate sheet

Re Item V
Reasoned statement under Article 35(2) with regard to novelty, inventive step or industrial applicability;
citations and explanations supporting such statement

1.   Reference is made to the following document:

D1:   WO 98 34357 A (NOKIA TELECOMMUNICATIONS OY ;PIIRAINEN OLLI
(FI)) 6 August 1998 (1998-08-06)

2.   The document D1 is regarded as being the closest prior art to the subject-matter
of claim 1, and shows (the references in parentheses applying to this document;
see especially Figure 2 and page 7, line 10 to page 13, line 13):

A system for generating a Doppler correction factor for compensating for Doppler
shift in a signal transmitted between a mobile station and a base station in a
mobile communication system, the system comprising:
a channel impulse response determination circuit (10) for determining a channel
impulse response for the channel on which the signal is received;
reference circuitry (14, 34) for generating a reference vector (ref) from a set of
known signals and the channel impulse response;
means (36) for generating an error vector (Ph-diff) from the reference vector (ref)
and received samples of the signal (r);
means (16) for determining noise variance (σ^2) of the incoming signal;
an estimator (18, 38 or 33) for generating a Doppler correction factor, the
estimator utilising the noise variance, the error vector (Ph-diff), the samples of the
received signal (r) and the sampling time (h_time).

The features of claims 1 and 9 which are not known from document D1 relate to
the Doppler correction factor having a real part representing amplitude correction
and an imaginary part representing phase correction. In document D1 (see e.g.
equations 6, 7 and 9), only the phase change resulting from the Doppler effect is
calculated and corrected.

Therefore, the subject-matter of claims 1 and 9 is new (Article 33(2) PCT).

3.   It also involves an inventive step as required by Article 33(3) PCT, because none
of the available prior art documents suggests to take into account the amplitude
variation in the signal due to the Doppler effect.

Re Item VII
Certain defects in the international application

1.   The features of the claims are not provided with reference signs placed in
parentheses (Rule 6.2(b) PCT).

2.   Contrary to the requirements of Rule 5.1(a)(ii) PCT, the relevant background art
disclosed in the document D1 is not mentioned in the description, nor is this
document identified therein.

PATENT COOPERATION TREATY

PCT

INTERNATIONAL SEARCH REPORT

(PCT Article 18 and Rules 43 and 44)

Applicant's or agent's file reference 89350/VRD	**FOR FURTHER ACTION**	see Notification of Transmittal of International Search Report (Form PCT/ISA/220) as well as, where applicable, item 5 below.
International application No. PCT/EP 99/ 01618	International filing date *(day/month/year)* 10/03/1999	(Earliest) Priority Date *(day/month/year)*

Applicant NOKIA NETWORKS OY et.al.

This International Search Report has been prepared by this International Searching Authority and is transmitted to the applicant according to Article 18. A copy is being transmitted to the International Bureau.

This International Search Report consists of a total of ____2____ sheets.

[X] It is also accompanied by a copy of each prior art document cited in this report.

1. **Basis of the report**

   a. With regard to the **language,** the international search was carried out on the basis of the international application in the language in which it was filed, unless otherwise indicated under this item.

   [ ] the international search was carried out on the basis of a translation of the international application furnished to this Authority (Rule 23.1(b)).

   b. With regard to any **nucleotide and/or amino acid sequence** disclosed in the international application, the international search was carried out on the basis of the sequence listing :

   [ ] contained in the international application in written form.

   [ ] filed together with the international application in computer readable form.

   [ ] furnished subsequently to this Authority in written form.

   [ ] furnished subsequently to this Authority in computer readble form.

   [ ] the statement that the subsequently furnished written sequence listing does not go beyond the disclosure in the international application as filed has been furnished.

   [ ] the statement that the information recorded in computer readable form is identical to the written sequence listing has been furnished

2. [ ] **Certain claims were found unsearchable** (See Box I).

3. [ ] **Unity of invention is lacking** (see Box II).

4. With regard to the **title,**

   [X] the text is approved as submitted by the applicant.

   [ ] the text has been established by this Authority to read as follows:

5. With regard to the **abstract,**

   [X] the text is approved as submitted by the applicant.

   [ ] the text has been established, according to Rule 38.2(b), by this Authority as it appears in Box III. The applicant may, within one month from the date of mailing of this international search report, submit comments to this Authority.

6. The figure of the **drawings** to be published with the abstract is Figure No. ____2____

   [ ] as suggested by the applicant.

   [ ] None of the figures.

   [X] because the applicant failed to suggest a figure.

   [ ] because this figure better characterizes the invention.

Form PCT/ISA/210 (first sheet) (July 1998)

INTERNATIONAL SEARCH REPORT

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7   H04B7/01

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

IPC 7   H04B

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	WO 98 34357 A (NOKIA TELECOMMUNICATIONS OY ;PIIRAINEN OLLI (FI)) 6 August 1998 (1998-08-06)	1,4,5,9, 10
A	page 7, line 10 –page 13, line 13; figure 2	2,8,11
A	EP 0 534 399 A (AEG MOBILE COMMUNICATION) 31 March 1993 (1993-03-31) page 2, line 57 –page 4, line 58; figure 2	1,5,9,10
A	EP 0 731 587 A (AT & T CORP) 11 September 1996 (1996-09-11) column 5, line 3 –column 8, line 42; figure 1	1,9

☐ Further documents are listed in the continuation of box C.

☒ Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
28 January 2000	11/02/2000

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL – 2280 HV Rijswijk Tel. (+31–70) 340–2040, Tx. 31 651 epo nl, Fax: (+31–70) 340–3016	Burghardt, G

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Patent document cited in search report		Publication date	Patent family member(s)		Publication date
WO 9834357	A	06-08-1998	AU	1600097 A	25-08-1998
			EP	0958666 A	24-11-1999
			NO	993682 A	03-08-1999
EP 0534399	A	31-03-1993	DE	4132200 A	01-04-1993
			DE	59207709 D	30-01-1997
EP 0731587	A	11-09-1996	US	5729558 A	17-03-1998
			JP	8321854 A	03-12-1996
			US	5872801 A	16-02-1999

COMPENSATION OF DOPPLER SHIFT
IN A MOBILE COMMUNICATION SYSTEM.

The present invention relates to compensation of Doppler shift
in a mobile communication station.

In a mobile communication system, signals which are transmitted
from mobile stations moving relative to a base station are
subject to the well-known Doppler effect, which causes a
frequency shift in the frequency received at the base station
relative to that which was transmitted by the mobile station.
This frequency shift is referred to herein as the Doppler shift.
The Doppler shift is dependent upon the speed and direction of
movement of the mobile station relative to the base station.
Thus, the Doppler effect can provide an increase or a decrease
in the frequency, depending on the direction of movement of the
mobile station relative to the base station. The magnitude of
the Doppler shift is dependent on the speed with which the mobile
station is moving relative to the base station.

Existing mobile communication installations provide a form of
Doppler compensation, in that the frequency detection circuitry
within the base station which selects a particular signal on a
particular channel can take into account a certain amount of
Doppler shift in the signal.

AU 664626 relates to a method and circuit arrangement for
compensating for the Doppler shift in a radio signal propagating
between a base station and a mobile station when the mobile
station approaches and moves past the base station. As a mobile
station approaches the base station with a decreasing propagating
time delay, the propagating time delay is integrated at intervals
to determine how the propagating time delay varies with time.
This information is used to effect a change in the radio signal
frequency at a particular time to compensate for the sudden
Doppler shift as the mobile station moves past the base station.
In a TDMA system, the interrogation occurs at intervals equal to

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an integral number of time frames. In the GSM standard a time
frame comprises eight consecutive time slots and a single
transmission burst passes between a particular mobile station and
base station in any one time frame.

Thus, in this method of Doppler compensation decisions are made
in response to past and incoming signals to improve the reception
at a future time, i.e. it is a reactive system. It would be
desirable to implement a system which can actively compensate for
Doppler shifts in an incoming signal in real time.

According to one aspect of the invention there is provided a
method for compensating for Doppler shift in a signal transmitted
between a mobile station and a base station in a mobile
communication system, the method comprising:
    determining a channel impulse response for the channel on
which the signal is received;
    using the channel impulse response to estimate data bits of
a selected portion of the received signal;
    generating a reference vector using the channel impulse
response and the estimated data bits;
    determining a Doppler characteristic using the selected
portion of the received signal and the reference vector; and
    using the Doppler characteristic to provide a Doppler shift
compensation for the received signal.

According to another aspect of the invention there is provided
a system for compensating for Doppler shift in a signal
transmitted between a mobile station and a base station in a
mobile communication system, the system comprising:
    a channel impulse response determination circuit for
determining a channel impulse response for the channel on which
the signal is received;
    an estimation circuit connected to receive the received
signal on the channel impulse response and to estimate data bits
of a selected portion of the received signal using the channel
impulse response;

a reference generator for generating a reference vector using the channel impulse response and the estimated data bits;

circuitry for determining a Doppler characteristic using the selected portion of the received signal and the reference vector; and

a Doppler shift compensation circuit operable to use the Doppler characteristic to provide a Doppler shift compensation for the received signal.

The invention is particularly applicable in TDMA mobile communication systems where the signal comprises a transmission burst. The selected portion is located in the transmission burst close to a zero phase offset point so that the effect of the Doppler characteristic is sufficiently small that it does not corrupt the transmitted bits.

According to the GSM standard, a TDMA transmission burst comprises a training sequence which is normally used to determine the channel impulse response for the channel on which the signal is received. This is done by a convolution of the received training sequence with a stored version of the training sequence.

In existing systems, the channel impulse response is used to remove from the received signal the effects of the transmission channel on the signal, in particular multi-path and attenuation effects.

According to the GSM standard, the "cleaned up" and filtered signal is demodulated to remove the IQ modulation by means of which the data in the signal was transmitted. Then, the demodulated signal can be decoded to generate hard bits. This can be done by a Viterbi technique.

The Doppler shift is estimated from samples of the received signal. The estimate for Doppler shift is thus dependent on channel quality, and typically gets worse when the channel quality is poor and better as the channel quality improves. When

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there is no Doppler shift and the channel conditions are near the
sensitivity level of a receiver, the application of a Doppler
compensation algorithm degrades the performance of the receiver.

On the other hand, if there is a Doppler shift in poor channel
conditions near the sensitivity level of the receiver, the
receiver is not able to meet the reference sensitivity limits if
a Doppler compensation algorithm is implemented.

According to one embodiment of the present invention the method
for compensating for Doppler shift in a signal transmitted
between a mobile station and a base station in a mobile
communication system, includes:
        detecting the quality of the received signal; and
        implementing a Doppler shift compensation in dependence on
the detected signal quality.

Thus in this embodiment Doppler shift compensation is used only
or mainly in good enough channel conditions. This provides an
increase in performance for the receiver. The technique is
provided for ensuring that Doppler compensation does not degrade
the sensitivity of the receiver.

The step of detecting signal quality can include estimating the
noise energy component of the signal. This can either be used
itself to generate a Doppler correction modification factor for
controlling the Doppler shift compensation in dependence on the
detected signal quality, or to generate a signal to noise ratio
for the received signal, which would then be used to generate the
Doppler correction modification factor.

The Doppler compensation can be implemented as any appropriate
user defined function of signal quality. For example, it could
be a linear function or a step function.

In the described embodiment, the adaptive method for Doppler
correction applied to a GSM system first estimates the quality

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of the channel and uses the resulting modification factor to scale a calculated phase difference between a reference signal and the actual received signal.

For a better understanding of the present invention and to show how the same may be carried into effect, reference will now be made by way of example to the accompanying drawings in which:

Figure 1 is a diagram of a signal burst in a mobile communication system;

Figure 2 is a block diagram of circuitry for implementing modified Doppler shift compensation; and

Figure 2a is a block diagram of a revised Doppler correction modification factor generator circuit.

Figure 1 illustrates a normal burst in a mobile communication system according to the GSM standard. This figure represents a burst received at a base station. For a TDMA system according to the GSM standard, mobile stations transmit bursts as modulated signals on frequency channels allocated by a base station controller. One frequency channel may support up to eight bursts, each burst associated with a respective call, where each call is allocated a time slot in which to send the burst. Further details of a TDMA system according to the GSM standard are not described herein because they are known to a person skilled in the art.

The normal burst contains two packets of 58 bits (DATA) surrounding a training sequence (TRS) of 26 bits. Three tail bits (TS) are added at each end of the normal burst. The training sequence (TRS) is a predetermined sequence of bits which is sent by the mobile station (MS) and is known at the base station controller (BSC). It is utilised at the base station controller to estimate the impulse response of the channel over which the burst is sent. The actual information which is transmitted is located in the data bits (DATA) of the burst.

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As explained earlier, the environment through which a signal passes from a mobile station to a base station can vary considerably, depending, amongst other things, on the distance between the mobile station and the base station, and interference caused by buildings and other structures in the area. As a result, the signal strength and signal quality of the signal received at the base station varies widely. Moreover, for moving mobile stations, the signal received by the base station is subject to a Doppler shift which should be corrected.

The circuit described herein provide a Doppler shift correction. The particular circuit described in relation to Figure 2 provides a correction only in situations where the channel conditions are good enough to give adequate signal quality received at the base station. Thus, a modification factor S_c is generated dependent on signal quality, which is used to control the Doppler shift correction so that Doppler shift correction is applied only in good enough channel conditions. It should be appreciated that although the use of a modification factor S_c in advantageous it is not essential to implementing a Doppler correction *per se*.

Figure 2 illustrates a circuit 1 suitable for implementing a Doppler compensation in a GSM system. It should be understood that the various blocks in Figure 2, although illustrated as separate interconnected entities, do not necessarily represent separate physical entities, but are intended to represent diagrammatically the various steps which are carried out. The blocks could be implemented as circuits or a suitably programmed microprocessor may effect each of the functions which is individually assigned to the blocks.

An antenna 20 receives signals 11 from the mobile stations. The antenna 20 is connected, via an interconnect 21, to RF circuitry 22. This circuitry 22 operates on the received burst to downshift the frequency to the baseband frequency and to sample the burst to provide from the analogue signal digital sampled values. The output of RF circuitry 22 is a sampled burst r (in

digital form), sampled at the expected bit rate of the
transmitted signal. Figure 1 illustrates such a burst. The
output of circuitry 22 is supplied along line 24 to a channel
impulse response (C.I.R.) block 10, to a variance calculator 16
to enable estimation of the quality of the communication channel
(as described later), to filtering and equalisation circuitry 12,
to a phase difference calculator 36 and to transforming circuitry
40 to enable the estimation and application of a Doppler shift
correction to the burst r.

The top part of Figure 2 illustrates the circuitry required for
implementing the adaptive part of the system, to generate the
Doppler correction modification factor S_c. A memory 32 holds the
training sequence TRSref which is the predetermined sequence of
bits which is sent by the mobile station MS as a training
sequence and received at the base station as TRS_received. The
reference training sequence TRSref is supplied to a reference
generator 14 and to the channel impulse response (C.I.R.) block
10. The reference generator 14 also receives the estimated
channel impulse response h from the channel impulse response
block 10.

The C.I.R. block 10 receives the burst r, including the received
training sequence TRS_received and calculates an estimated
channel impulse response h by calculating the cross correlation
between the received training sequence TRSreceived and the known
training sequence TRSref.So,

        h = xcorr (TRS_received, TRSref)   (equation 1)

It will be appreciated that, prior to effecting the cross
correlation, the known training sequence TRSref, which is stored
in digital form, is i,q modulated similarly to the manner in
which the training sequence has been modulated at the MS for
transmission, according to the GSM standard. The cross
correlation is done in a known manner to produce a channel
impulse response in the form of five tap values $(h(i)_{i=0 to 4})$.

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As is known, the estimated impulse response h is used to calculate the expected estimate of the data in the received burst, as though the data has been subject to the same average noise.

The C.I.R. block also generates timing advance information τ, which is used to determine where in the allocated time slot the received burst r is located.

For each burst, the estimated channel impulse response h for that burst is calculated by the CIR block 10 and is supplied to filtering/equalising circuitry which allows the data, DATA(r), in that burst to be recovered. As is known, the filtering/equalising circuit 12 receives the channel impulse response h and timing information τ for the received burst to allow the signal to be demodulated, filtered and decoded, to recover the data in a known manner.

The reference generator 14 produces a reference vector, reffi, which is calculated using the convolution of the impulse response and the known training sequence. Thus, the reference generator 14 performs the following calculation:

$$reffi = h*TRSref \qquad\qquad\qquad\qquad (equation\ 2)$$

In more detail, (where $reffi_k$ represents the kth sample of the signal reffi)

$$reffi_k \;=\; \sum_{i=0}^{N-1} h_i \cdot (1-2.TRS_{k-i}) \qquad\qquad (equation\ 3)$$

in which N represents the number of tap values in the estimated impulse response h (N = 5 in the described embodiment), and k runs from N-1 to 25.

The vector reffi is supplied from the reference generator to the

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variance calculator 16. As described above, the variance
calculator also receives the burst r, including the received
training sequence. The variance calculator calculates a variance
var (σ^2) according to the following equation:

$$var = \frac{\sum_{k=4}^{25}(|r_k - reffi_k|^2))}{reffi\_length} \qquad (equation\ 4)$$

The term reffi_length is a constant representing the length of
the reference signal, reffi. This is calculated by multiplying
the number of samples (22) by the bit separation.

In equation 4, the values of r_k are the sampled values of the
received training sequence for the burst r.

It will be appreciated that each actual received sample r_k will
have a noise level which is different to the averaged estimated
noise level derived from the channel impulse response and
reflected in the reference samples $reffi_k$. Thus, the variance
gives an indication of the level of noise energy actually
received, and thus signal quality.

The output σ^2 of the variance calculator 16 is supplied to a
Doppler correction modification factor circuit 18. The Doppler
correction modification factor circuit 18 uses the calculated
variance σ^2 to generate a modification factor S_c using a function
which can be determined by a user. In the embodiment shown in
Figure 2, the Doppler correction modification factor circuit
generates the modification S_c as a function of the variance σ^2,
for example a linear function or a non-linear function such as
a step function.

In another embodiment of the invention the value of S_c is
calculated in dependence upon a signal to noise ratio (SNR) of
the channel. Figure 2a illustrates circuitry for a Doppler

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correction modification circuit 18' for implementing such an embodiment. A SNR calculator 42 receives the calculated variance σ^2 from the variance calculator 16 and an energy value E. The signal energy E may be determined from the calculated signal reffi in accordance with the following:

$$E = \sum_{k=4}^{25} \frac{|reffi_k|^2}{reffi\_length} \qquad \text{(equation 4a)}$$

Alternatively the tap values can be used to derive the signal energy, E, of the channel according to the following:

$$E = \sum_{i=0}^{4}(h(i))^2$$

The SNR value is calculated by the SNR calculator 42 in accordance with the following equation:

$$SNR = \frac{E}{\sigma^2} \qquad \text{(equation 4b)}$$

The above referenced technique is described in our earlier Application No. PCT/FI96/00461, the contents of which are herein incorporated by reference.

The SNR value is then supplied to a revised modification factor generating circuit 44. This circuit 44 calculates the value of the modification factor S_c as a function of the SNR value, for example a linear function or a non-linear function such as a step function.

The lower part of the circuit in Figure 2 illustrates in block diagram form a system for implementing a Doppler shift

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correction. However, it will be apparent that the Doppler
correction adaptive part described above could be used with other
implementations of Doppler correction.

The main function of the adaptive part of the circuit described
above is to provide a Doppler correction modification factor S_c
which is based on the channel conditions, so that Doppler
correction is used only or mainly in good enough channel
conditions. This is useful in any type of Doppler correction.
A particular Doppler correction circuit is discussed in the
following.

An equalisation circuit 30, for example a Viterbi equaliser,
receives the filtered, demodulated and equalised signal DATA(r)
from the filtering and equalisation circuitry 12. The
equalisation circuit 30 operates on a part of the data sequence
DATA of the burst (that part having been derived from ESTIM.BLOCK
in Figure 1) to estimate and output bits which were sent from the
mobile station MS. This output is referred to herein as
estim_bits, and they run from k=j to k=j+n. The equalisation
circuit 30 operates to make decisions of the bits as in known
mobile communication systems and thus it will not be described
further herein.

The estimated bit decisions estim bits are supplied to a
reference circuit 34. The reference circuit 34 generates a
reference vector ref by using a convolution of the estimated bit
decisions and the estimated impulse response h, according to the
following equation:

$$ref = estim\_bits * h \qquad\qquad\qquad (equation\ 5)$$

Thus, the reference vector ref comprises a set of samples ref_k
$(k=j\rightarrow j+n)$, each having real and imaginary values. The reference
vector ref is supplied to a phase difference calculator 36. As
described earlier, the phase difference calculator 36 also

receives the received burst r. As known in the art, the received burst comprises samples r_k each having real and imaginary values.

The phase difference calculator uses a value t_diff which represents the time between a zero phase offset point h_time and the middle of the estimation block, as illustrated in Figure 1. The zero phase offset point h_time is a zero phase offset point inside the training sequence where the calculated impulse response is true. In practice, this is typically the middle of the training sequence. The value of t_diff and h_time can be determined during the design phase of the system, and a constant thereafter. Of course, they could be reprogrammed if necessary during use of the system.

In addition, the location (j) of the beginning of the estimation block estim_block and its length (n) is determined and programmed into the equalisation circuit 30. The estimation block is selected so that the Doppler offset has not yet corrupted the received bits.

The phase change per bit duration (ph_diff) resulting from the Doppler effect (the Doppler characteristic) is calculated from the reference signal ref and the actual received signal r by the phase difference calculator 36 according to one of the following equations:

$$ph\_diff = \frac{1}{t\_diff} \tan^{-1}\left\{\frac{\sum_k imag(r_k \cdot ref*_k)}{\sum_k real(r_k \cdot ref*_k)}\right\} \qquad \text{(equation 6)}$$

$$ph\_diff = \frac{1}{t\_diff} \sum_k \cdot \tan^{-1}\left\{\frac{imag(r_k \cdot ref*_k)}{real(r_k \cdot ref*_k)}\right\}/length(k)$$

$$\text{(equation 7)}$$

where k runs from j to j+n, and where length (k) represents the amount of different k values in the summation, i.e. n.

A Doppler correction circuit 38 is then used to correct the

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estimated Doppler shift from the received samples. The Doppler correction circuit 38 receives the zero phase offset point h_time and the Doppler correction modification factor S_c. Furthermore, it receives the calculated phase difference from the phase difference calculator 36. Knowing that the point h_time has zero phase offset, the actual Doppler phase shift ϕ can be calculated for each bit as follows:

$$\phi_k = S_c \cdot ph\_diff \cdot (k-h\_time) \qquad \text{(equation 8)}$$

where k is a bit index of the received sample r. When the index k<h_time the phase shift has an opposite sign to when k>h_time. It will be apparent that if the Doppler correction circuitry is operating in independence of the adaptive circuitry which produces Sc, Sc will default to 1 in equation 8.

Transforming circuitry 40 is then used to implement the Doppler shift correction on the received burst r to produce a corrected signal. The transforming circuitry receives the estimated Doppler shift vector ϕ (comprising the ϕ_k values) and sampled values of the received burst r. It performs a CORDIC operation to correct for the Doppler shift of each sample, according to the following operation.

$$\begin{pmatrix} N\_real\_sample(k) \\ N\_imag\_sample(k) \end{pmatrix} = \begin{pmatrix} \cos\phi_k & \sin\phi_k \\ \sin\phi_k & \cos\phi_k \end{pmatrix} \begin{pmatrix} real(r_k) \\ imag(r_k) \end{pmatrix}$$

$$\text{(equation 9)}$$

The Doppler shift corrected vector DCV which is output from the transforming circuitry 40 is supplied to the filtering/equalising circuit 12, so that the Doppler corrected signal is used to recover the data from the signal.

As part of the Doppler correction technique described above, bit decisions for the estimation block have already been made. These constitute part of the data. It is thus not necessary to estimate the same bits again, although this could be done.

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Instead, the equalisation circuit 30 can be stopped at the end
of the estimation block and the current state preserved. Next,
the Doppler correction is performed for the remaining bits and
the Viterbi equalisation is then executed to the end of the time
slot on the Doppler corrected bits. In the second part, the
Doppler correction may be done to the first part of the time slot
and the Viterbi estimation can then be performed for the data 58
in the first part of the time slot. This method reduces the
required calculations in the receiver.

It is possible to implement a limit on the value of the phase
difference ph_diff, so that if the phase difference is below a
certain threshold, no correction is performed.

A typical environment where Doppler correction could be used is
fast trains or motorways. In that situation, it is likely that
there would be a line of sight path from the base station to the
mobile station, so that the Doppler correction would be a
constant value between different time slots if the velocity of
the mobile station is the same. In that case, an average value
of the phase difference could be calculated from several
different time slots, and this average value could be used as a
correction value.

In the case of a diversity receiver having a plurality of
different branches, the phase differences can be calculated for
all branches using the same estimates as the received samples,
with each branch using its own impulse response.

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CLAIMS:

1.   A method for compensating for Doppler shift in a signal transmitted between a mobile station and a base station in a mobile communication system, the method comprising:
determining a channel impulse response for the channel on which the signal is received;
using the channel impulse response to estimate data bits of a selected portion of the received signal;
generating a reference vector using the channel impulse response and the estimated data bits;
determining a Doppler characteristic using the selected portion of the received signal and the reference vector; and
using the Doppler characteristic to provide a Doppler shift compensation for the received signal.

2.   A method according to claim 1, wherein the signal comprises a transmission burst in a TDMA mobile communication system.

3.   A method according to claim 1, wherein the selected portion is located in said transmission burst close to a zero phase offset point, whereby it has been substantially unaffected by the Doppler characteristic.

4.   A method according to claim 2 or 3, wherein the channel impulse response is determined from a training sequence in said transmission burst.

5.   A method according to claim 1, 2, 3 or 4, wherein the data bits are estimated by:
using the channel impulse response to remove from the received signal the effects of the transmission channel for the signal;
demodulating the resulting signal; and
decoding the selected portion of the demodulated signal to estimate data bits.

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6.   A method according to claim 5, wherein the decoding step is done by a Viterbi method.

7.   A method according to any preceding claim, wherein the step of using the Doppler characteristic comprises determining for each of a plurality of indexed samples of the received signal the Doppler phase shift for that sample based on the Doppler characteristic and the location of the induced sample within the received signal.

8.   A method according to claim 7, where the samples are indexed at the bit rate of the received signal.

9.   A system for compensating for Doppler shift in a signal transmitted between a mobile station and a base station in a mobile communication system, the system comprising:
     a channel impulse response determination circuit for determining a channel impulse response for the channel on which the signal is received;
     an estimation circuit connected to receive the received signal on the channel impulse response and to estimate data bits of a selected portion of the received signal using the channel impulse response;
     a reference generator for generating a reference vector using the channel impulse response and the estimated data bits;
     circuitry for determining a Doppler characteristic using the selected portion of the received signal and the reference vector; and
     a Doppler shift compensation circuit operable to use the Doppler characteristic to provide a Doppler shift compensation for the received signal.

10.   A system according to claim 9, wherein the estimation circuit comprises filtering an equalisation circuit operable to remove from the received signal the effects of the transmission channel for the signal by using the channel impulse response;
     demodulation circuitry for demodulating the resulting

signal; and

decoding circuitry for decoding the selected portion of the demodulated signal to estimate the data bits.

11. A system according to claim 10, wherein the decoding circuitry is a Viterbi decoder.

AMENDED CLAIMS
[received by the International Bureau on 3 March 1998 (3.03.98);
original claims 1,2,3,9 and 10 amended; remaining
claims unchanged (3 pages)]

1.   A method for compensating for Doppler shift in a signal
transmitted between a mobile station and a base station in a
mobile communication system, said signal comprising a sequence
of transmission bursts, the method comprising:

determining for each transmission burst a channel impulse
response for the channel on which the signal is received;

using the channel impulse response to estimate data bits of
a selected portion of the transmission burst; said selected
portion being located in said transmission burst close to a zero
phase offset point;

generating a reference vector using the channel impulse
response and the estimated data bits;

determining a Doppler characteristic in the form of a phase
shift per bit duration using the selected portion of the received
signal and the reference vector; and

using the Doppler characteristic determined for the selected
portion to provide a Doppler shift compensation for the
transmission burst.

2.   A method according to claim 1, when used in a TDMA mobile
communication system.

3.   A method according to claim 1, wherein the Doppler shift
compensation ϕ_k for each bit is calculated as follows, wherein k
is the bit index, h_time is the point of zero phase offset and
ph_diff is the phase shift per bit duration,

$$ph\_diff \cdot (k-h\_time).$$

4.   A method according to claim 2 or 3, wherein the channel
impulse response is determined from a training sequence in said
transmission burst.

5.   A method according to claim 1, 2, 3 or 4, wherein the data
bits are estimated by:

using the channel impulse response to remove from the
received signal the effects of the transmission channel for the
signal;

demodulating the resulting signal; and

decoding the selected portion of the demodulated signal to
estimate data bits.

6.   A method according to claim 5, wherein the decoding step is
done by a Viterbi method.

7.   A method according to any preceding claim, wherein the step
of using the Doppler characteristic comprises determining for
each of a plurality of indexed samples of the received signal the
Doppler phase shift for that sample based on the Doppler
characteristic and the location of the induced sample within the
received signal.

8.   A method according to claim 7, where the samples are indexed
at the bit rate of the received signal.

9.   A system for compensating for Doppler shift in a signal
transmitted between a mobile station and a base station in a
mobile communication system, said signal comprising a sequence
of transmission bursts, the system comprising:

a channel impulse response determination circuit for
determining for each transmission burst a channel impulse
response for the channel on which the signal is received;

an estimation circuit connected to receive the received
signal on the channel impulse response and to estimate data bits
of a selected portion of the transmitted burst using the channel
impulse response, said selected portion being located in said
transmission burst close to a zero phase offset point;

a reference generator for generating a reference vector
using the channel impulse response and the estimated data bits;

circuitry for determining a Doppler characteristic in the
form of a phase change per bit duration using the selected
portion of the received signal and the reference vector; and

a Doppler shift compensation circuit operable to use the
Doppler characteristic to provide a Doppler shift compensation
for the transmission burst.

10. A system according to claim 9, wherein the estimation
circuit comprises:

filtering and equalisation circuitry operable to remove from
the received signal the effects of the transmission channel for
the signal by using the channel impulse response;

demodulation circuitry for demodulating the resulting
signal; and

decoding circuitry for decoding the selected portion of the
demodulated signal to estimate the data bits.

11. A system according to claim 10, wherein the decoding
circuitry is a Viterbi decoder.

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FIG. 1

FIG. 2a

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FIG. 2

(54) Zeitmultiplex-Verfahren zur Bestimmung der mittleren Phasenänderung eines Empfangssignals.

(57) Die Erfindung betrifft ein Zeitmultiplex-Verfahren, mit dem die mittlere Phasenänderung eines Empfangssignals bestimmt wird. Im Mobilfunk ist die Kanalimpulsantwort durch den Dopplereffekt zeitlich veränderlich. Es stellt sich das Problem, diese Änderungen in einem Empfänger zu schätzen und zu berücksichtigen. Es wird ein einfaches Verfahren angegeben, welches die Änderungen der Kanalimpulsantwort durch eine mittlere Phasenänderung beschreibt. Das neue Verfahren ist für TDMA Systeme wie zum Beispiel das GSM System geeignet.

Fig. 2

Die Erfindung betrifft ein Verfahren nach dem Oberbegriff des Patentanspruches 1.

Die Erfindung findet Verwendung bei digitalen Kommunikationsnetzen, insbesondere Mobilfunksystemen. Das Verfahren ist etwa für TDMA (Time Division Multiplexing Access) Systeme, z.B. das GSM (Group Spéciale Mobile) System geeignet.

5 Im Mobilfunk ist die Kanalimpulsantwort durch den Dopplereffekt zeitlich veränderlich. Es stellt sich das Problem, diese Änderungen in einem Empfänger zu schätzen und zu berücksichtigen.

Vorbekannte Lösungen sind beispielsweise:

a) die Kanalimpulsantwort wird im gesamten Zeitschlitz als konstant angenommen.

b) Die Kanalimpulsantwort wird vollständig im Zeitschlitz nachgeführt. Dies bedeutet, die zeitliche

10 Veränderung der Mehrwegeausbreitung und damit die Form der Kanalimpulsantwort für alle Zeitpunkte im Zeitschlitz zu schätzen. Die Nachführung kann mit Hilfe eines Kalmanfilters geschehen. In jedem Falle ist entweder eine Entscheidungsrüc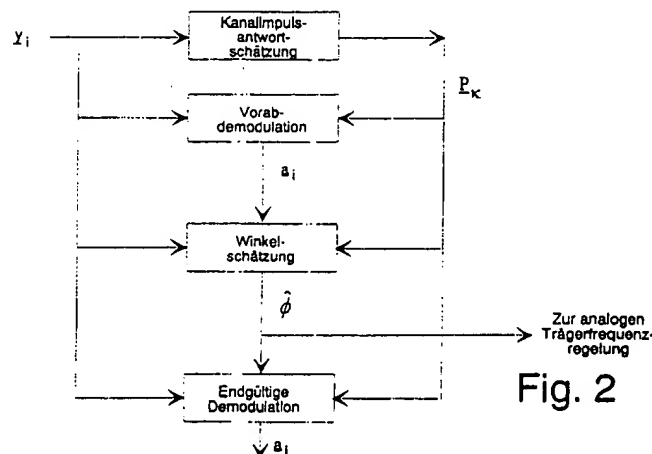kkopplung erforderlich oder es sind weitere Stützstellen (d.h. dem Empfänger bekannte feste Sequenzen) im Zeitschlitz erforderlich. Auch ist eine Interpolation der Kanalimpulsantwort im ganzen Zeitschlitz möglich, falls hintereinanderliegende Zeitschlitze auf der

15 gleichen Frequenz entsprechend dicht liegen und die Geschwindigkeit nicht zu groß ist (Literatur: Hoher, P.: Kohärenter Empfang trelliscodierter PSK-Signale auf frequenzselektiven Mobilfunkkanälen - Entzerrer, Decodierung und Kanalparameterschätzung, Fortschritt-Bericht VDI, Dissertation, Universität Kaiserslautern, Mai 1990).

Die bekannten Verfahren haben den Nachteil, daß bei hohen Geschwindigkeiten sich eine entsprechend

20 hohe Bitfehlerrate ergibt und daß entsprechende Verfahren sehr aufwendig sind.

Im Falle von Entscheidungsrückführung reagieren diese Verfahren sehr empfindlich auf Bitfehler. Bei kleiner Geschwindigkeit und hohen Störpegeln ergibt sich eine große Degradation der Empfindlichkeit im Vergleich zu einem einfachen Empfänger, bei dem die Kanalimpulsantwort im Zeitschlitz als konstant angenommen wird.

25 Werden Stützstellen im Zeitschlitz verwendet, so bedeutet dies natürlich eine Verkleinerung der Übertragungskapazität. Die Benutzung von Nachbarzeitschlitzen verbietet die Benutzung eines Frequenzsprungverfahrens, welches große Vorteile bei Fading aufweißt und die TDMA Struktur muß an die maximale Geschwindigkeit angepasst werden.

Der Erfindung liegt deshalb die Aufgabe zugrunde, ein Verfahren anzugeben, bei dem eine Dopplerkor-

30 rektur und Trägerfrequenzkorrektur bei Übertragung in Zeitmultiplex-Systemen mit linearen, digitalen Modulationen mit konstanter einhüllender unter Minimierung des Aufwandes bei gleichzeitig hoher Empfangsqualität erreicht wird. Diese Aufgabe wird gelöst durch die im kennzeichnenden Teil des Patentanspruch 1 angegebenen Merkmale. Vorteilhafte Ausgestaltungen und/oder Weiterbildungen sind den Unteransprüchen zu entnehmen.

35 Es wird davon ausgegangen, daß die Zeitschlitze des TDMA-Systems in der Mitte eine dem Empfänger bekannte Trainingssequenz enthalten, mit deren Hilfe die Kanalimpulsantwort gemessen werden kann. Die Erfindung besteht zunächst aus der Vorabdemodulation des mittleren Teiles eines Zeitschlitzes. Danach kann eine mittlere Phasendrehung berechnet werden. Diese beschreibt in guter Näherung auch bei Mehrwegeausbreitung und bei beliebigem Dopplerspektrum (z.B. symmetrisches Dopplerspektrum) die

40 Änderung der Kanalimpulsantwort innerhalb eines Zeitschlitzes. Die endgültige Demodulation erfolgt dann in mehreren Bereichen innerhalb eines Zeitschlitzes. In jedem Bereich wird hierzu die Kanalimpulsantwort als konstant angenommen und entsprechend der mittleren Phasendrehung berechnet. Nach Mittelung über mehrere Zeitschlitze kann die Phasendrehung auch für die Berechnung der Stellgröße für die analogen Trägerfrequenzerzeugung verwendet werden.

45 Die Erfindung hat den Vorteil, daß lediglich kleine Bitfehlerraten bei hohen Geschwindigkeiten auch bei Mehrwegeausbreitung mit einer maximalen Mehrwegeverzögerung, welche größer als die Symboldauer ist, auftreten. Gleichzeitig ist nur ein geringer Mehraufwand erforderlich im Vergleich zu einem einfachen Verfahren, welches von einer konstanten Kanalimpulsantwort über den gesamten Zeitschlitz ausgeht. Durch die Dopplerkorrektur ergibt sich keine Verschlechterung der Empfindlichkeit bei kleinen Geschwindigkeiten.

50 Weitere Stützstellen außerhalb der Trainingssequenz werden nicht benötigt. Bei Intersymbolinterferenz kann wie bei einem einfachen Verfahren z.B. ein Viterbi Entzerrer verwendet werden, der von einer konstanten Autokorrelationsfunktion der Kanalimpulsantwort über den gesamten Zeitschlitz ausgeht. Die Erfindung benötigt kein Gedächtnis über mehr als einen Zeitschlitz und damit ist ein Frequenzsprungverfahren möglich.

55 Die Erfindung wird im folgenden anhand von Ausführungsbeispielen beschrieben unter Bezugnahme auf schematische Zeichnungen.

Es wird ein digitales Modulationsverfahren (z.B. Phase Shift Keying PSK oder Minimum Shift Keying MSK) und ein linearer Übertragungskanal betrachtet. Es ist allgemein bekannt, daß das Empfangssignal y(t)

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dann durch die Faltung einer Kanalimpulsantwort p(t,τ) mit den gesendeten komplexen oder reellen Symbolen a_k beschrieben werden kann:

$$\underline{y}(t) = \int_{-\infty}^{\infty} \underline{p}(t,\tau) \sum_k \underline{a}_k \delta(t - \tau - kT) \mathrm{d}\tau$$

Die Kanalimpulsantwort gibt sowohl die Eigenschaften der Modulation wie des Funkkanales wieder (Mehrwegeausbreitung). Sie beschreibt die Antwort der Übertragungsstrecke zum Zeitpunkt t auf einen Dirac-Stoß als Eingangssignal in Abhängigkeit der Verzögerung τ. Die Kanalimpulsantwort wird zunächst als zeitveränderlich angesehen. Durch Dopplerverschiebungen ändert sich z.B. im Mobilfunk die Form der Kanalimpulsantwort nach einer gewissen Wegstrecke vollständig. Ist die maximale Mehrwegeverzögerung klein gegenüber der Symboldauer, so kann die zeitliche Veränderung der Kanalimpulsantwort durch eine Amplituden- und Phasenänderung beschrieben werden. Die Form der Kanalimpulsantwort hängt dann nicht mehr vom Funkkanal ab. Es wird nun ein TDMA System betrachtet, bei dem für die Mehrwegeverzögerung mehrere Symbollängen zugelassen werden. Jeder Zeitschlitz besitzt in der Mitte eine feste dem Empfänger bekannte Symbolsequenz (Trainingssequenz). Mit ihr wird die Kanalimpulsantwort in der Mitte des Zeit-schlitzes gemessen werden. Für eine richtige Demodulation muß jedoch auch die zeitliche Änderung der Kanalimpulsantwort innerhalb eines Zeitschlitzes bekannt sein. Auch dann, wenn die Mehrwegeverzögerung die Dauer eines Symboles deutlich überschreitet, wird die zeitliche Änderung der Kanalimpulsantwort immer noch durch eine mittlere Amplituden- und Phasenänderung beschrieben . Voraussetzung ist, daß der Zeitschlitz nicht zu lang ist, damit die Änderung der Kanalimpulsantwort nicht zu groß wird. Mit dieser Voraussetzung ist der Fehler der obigen Näherung gering. Es gibt also einen Wertebereich für die Länge eines Zeitschlitzes in dem die Annahme einer zeitinvarianten Kanalimpulsantwort zu merklichen Verlusten der Leistungsfähigkeit bei der Demodulation führt. Die Annahme einer mittleren konstanten Amplituden- und Phasenänderung der Kanalimpulsantwort innerhalb eines Zeitschlitzes führt dann zu einer deutlichen Verbesserung.

Bei Phasenmodulationsarten mit konstanter Einhüllender steckt die Information nur in der Phase und auch bei einer Änderung der Amplitude der Kanalimpulsantwort kann ohne Kenntnis dieser Amplitudenänderung noch richtig demoduliert werden. Deswegen reicht es dann hier aus, nur die mittlere Änderung der Phase zu betrachten.

Es wird angenommen, daß das Empfangssignal bandbegrenzt ist und daher abgetastet werden kann. Die Abtastrate soll n mal die Symbolrate betragen. Das abgetastete Empfangssignal wird nun durch die diskrete Faltung der Symbole mit der abgetasteten Kanalimpulsantwort beschrieben werden.

$$\underline{y}_{ni+k} = \sum_l \underline{a}_{(i-l)} \underline{p}_{t,nl+k} \qquad k = 0, \cdots, n-1$$

Y_{ni+k} sind die empfangenen komplexen Abtastwerte, $P_{t,nl+k}$ ist die Kanalimpulsantwort und die a_i sind die gesendeten komplexen (z.B. 4,8 PSK) oder reellen Symbole (z.B. 2PSK oder MSK).

Zunächst wird die Kanalimpulsantwort in der Mitte des Zeitschlitzes mit Hilfe einer festen dem Empfänger bekannten Sequenz (Trainingssequenz) geschätzt. Dann kann in einem mittleren Teil des Zeitschlitzes (siehe FIG. 1) mit Hilfe eines digitalen Matched Filters (an die gemessene Kanalimpulsantwort angepaßt) eine Symbolschätzung vorgenommen werden (Vorabdemodulation).

$$\mathrm{mfo}_i = \sum_k \underline{y}_{ni+k} \underline{p}_k^*$$

Die empfangenen Symbole lassen sich durch eine Harddecision des Ausgangs des Matched Filters mfo_i bestimmen. Es werden nun rechts und links der Trainingssequenz zwei Bereiche definiert, deren Symbole nun bekannt sind. Diese Bereiche werden im folgenden Ersatzpräambeln genannt. Es wird nun jeweils die mittlere Phasendrehung der Kanalimpulsantwort zwischen dem Bereich einer Ersatzpräambel und dem Bereich der Trainingssequenz geschätzt. Durch Mittelung beider Winkel wird ein endgültiger Wert für die

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Phasendrehung bestimmt.

Im folgenden wird nun die Schätzung der Phasendrehung mit Hilfe einer Ersatzpräambel und der bereits bekannten Kanalimpulsantwort beschrieben. Es wird vorausgesetzt, daß die Ersatzpräambel so kurz sein soll, daß man die Phase ϕ innerhalb der Ersatzpräambel als konstant ansetzen kann. Das Empfangssignal ist dann durch folgende Gleichung gegeben.

$$\underline{y}_{ni+k} = e^{j\phi} \sum_l \underline{a}_{i-l} \underline{p}_{nl+k}$$

Nach dem Maximum Likelihood Prinzip kann jetzt die Phase geschätzt werden. Der folgende Ausdruck muß dazu minimiert werden (der Rauschanteil aufeinanderfolgender Abtastwerte wird als unkorreliert angesehen, weißes Rauschen):

$$\sum_i \sum_{k=0}^{n-1} |\underline{y}_{ni+k} - e^{j \cdot \hat{\phi}} \sum_l \underline{a}_{i-l} \underline{p}_{nl+k}|^2$$

Die Minimierung erfolgt bezüglich $\hat{\phi}$. Alle anderen Größen sind bereits bekannt und damit fest. Die Minimierung des obigen Ausdrucks läßt sich dann durch die Maximierung des folgenden Ausdrucks erreichen:

$$\mathrm{Re}\{e^{-j \cdot \hat{\phi}} \sum_l \sum_{k=0}^{n-1} \underline{p}^*_{nl+k} \sum_i \underline{y}_{ni+k} \underline{a}^*_{i-l}\}$$

$$= \mathrm{Re}\{e^{-j \cdot \hat{\phi}} \sum_l \sum_{k=0}^{n-1} \underline{p}^*_{nl+k} \sum_i \underline{y}_{n(l+i)+k} \underline{a}^*_i\}$$

Der Realteil wird maximal, falls für $\hat{\phi}$ der Phasenwinkel der Summe gewählt wird. Dies bedeutet, daß zur Berechnung der Phase $\hat{\phi}$ die konjugiert komplexen vorabdemodulierten Symbole der Ersatzpräambel mit dem Empfangsabtastwerten korreliert werden müssen.

Dies entspricht einer erneuten Schätzung der Kanalimpulsantwort. Die bereits bekannte Kanalimpulsantwort wird nun mit der neu geschätzten konjugiert komplexen Kanalimpulsantwort Abtastwert für Abtastwert multipliziert und aufakkumuliert.

Da entsprechend den obigen Voraussetzungen die Phase über einen größeren Bereich als konstant angesehen werden kann, braucht auch bei der endgültigen Bestimmung der empfangenen Symbole der Phasenwinkel nicht Symbol für Symbol nachgeführt werden. Es reicht aus, den Zeitschlitz in mehrere Bereiche einzuteilen, in denen die Phase jeweils wieder als konstant angesehen werden kann (siehe FIG. 1). Die Symbole werden wieder mit Hilfe eines Matched Filters geschätzt. Die Kanalimpulsantwort wird für jeden Bereich entsprechend der geschätzten Drehung und dem Abstand von der Trainingssequenz bestimmt:

$$\underline{p}_{d,k} = \underline{p}_{0,k} \cdot e^{j\hat{\phi}d/\Delta}$$

Δ ist der Abstand einer Ersatzpräambel zur Trainingssequenz und d ist der mittlere Abstand eines Bereiches für das endgültige Matched Filter zur Trainingssequenz. Zur Korrektur der Intersymbolinterferenz können bekannte Standardverfahren angewendet werden (z.B. Viterbi Entzerrer). Es muß hierbei nicht mehr zwischen verschiedenen Bereichen im Zeitschlitz unterschieden werden, da die Autokorrelationsfunktion der Kanalimpulsantwort nicht von einer Phasendrehung abhängig ist.

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FIG. 2 zeigt ein Blockschaltbild des vollständigen Empfängers.

In jedem Empfänger besteht das Problem die Trägerfrequenz analog nachzuführen. Dies kann mit dem obigen Verfahren kombiniert werden. Die geschätzte Phasendrehung gibt nicht nur die Dopplerverschiebung wieder, sondern nach entsprechender Mittelung auch den Versatz der Trägerfrequenz (siehe FIG. 3).

Im folgenden werden einige Simulationsergebnisse für einen GSM Empfänger nach obigem Verfahren gegeben. Das endgültige Matched Filter arbeitet hierbei in drei verschiedenen Bereichen und als Modulation wird GMSK verwendet (binäre Modulation). Zum Vergleich werden Ergebnisse für einen Empfänger dargestellt, der eine konstante Kanalimpulsantwort im gesamten Zeitschlitz annimmt. Als Ausbreitungsmodell wird das sogenannte "Rural Area" Profil nach den COST 207 Spezifikationen verwendet. Dieses Profil ist im GSM System eines der Referenzprofile zur Spezifikation der erlaubten Bitfehlerraten. Angegeben wird im folgenden die Restfehlerwahrscheinlichkeit des Demodulators bei Empfang eines Signales mit dem obigen Mehrwegeprofil ohne weitere Störungen durch Rauschen in Abhängigkeit der Geschwindigkeit V.

V in km/h	50	250	500
Neues Verfahren	$3.2 \cdot 10^{-4}$	$5.6 \cdot 10^{-3}$	$2.1 \cdot 10^{-2}$
Verfahren mit konst. angenommener Kanalimpulsantwort	$3.8 \cdot 10^{-4}$	$9.0 \cdot 10^{-3}$	$3.3 \cdot 10^{-2}$

Die Verbesserung durch das neue Verfahren insbesondere bei hohen Geschwindigkeiten ist deutlich zu sehen.

Die folgende Tabelle gibt die Bitfehlerrate des Demodulators bei Empfang des obigen Profiles bei 50 km/h und mit Rauschen (E_b/N_o = 6 dB) an:

Neues Verfahren	$6.8 \cdot 10^{-2}$
Verfahren mit konst. angenommener Kanalimpulsantwort	$6.8 \cdot 10^{-2}$

Daraus geht deutlich hervor, daß sich durch die Dopplerkorrektur bei kleinen Geschwindigkeiten keine Verschlechterung der Empfindlichkeit ergibt.

Patentansprüche

1. Zeitmultiplex-Verfahren zur Bestimmung der mittleren Phasenänderung des Empfangssignals, bei dem für die Mehrwegeverzögerung mehrere Symbollängen verwendet werden und jeder Zeitschlitz in der Mitte eine feste dem Empfänger bekannte Symbolsequenz, sog. Trainingssequenz, besitzt, dadurch gekennzeichnet,
   - daß die Kanalimpulsantwort innerhalb eines Zeitschlitzes mit einer festen dem Empfänger bekannten Trainingssequenz geschätzt wird,
   - daß im mittleren Teil des Zeitschlitzes eine an die geschätzte Kanalimpulsantwort angepaßte Symbolschätzung durch Demodulation vorgenommen wird,
   - daß mit den empfangenen Symbolen rechts und links der Trainingssequenz zwei Bereiche definiert werden, sog. Ersatzpräambeln,
   - daß zwischen dem Bereich der Ersatzpräambeln und dem Bereich der Trainingssequenz die mittleren Phasendrehungen der Kanalimpulsantwort geschätzt werden und durch Mittelung über die Winkel der Schätzwert für die genaue Phasendrehung verbessert wird.

2. Verfahren nach Anspruch 1, dadurch gekennzeichnet,
   - daß die Phasendrehung der Kanalimpulsantwort mit einem Empfangssignal mit konstanter Phase innerhalb der Ersatzpräambel dadurch bestimmt wird, daß die konjugiert komplexen, vorab demodulierten Symbole der Ersatzpräambel mit den Empfangssignalabtastwerten korreliert werden und die Kanalimpulsantwort erneut geschätzt wird, und
   - daß die bekannte Kanalimpulsantwort mit der neu geschätzten, konjugiert komplexen Kanalimpulsantwort Abtastwert für Abtastwert multipliziert und aufakkumuliert sind.

3. Verfahren nach Anspruch 2, dadurch gekennzeichnet, daß der Zeitschlitz in mehrere kurze Bereiche annähernd konstanter Phase eingeteilt wird,
   - daß die Kanalimpulsantwort für jeden Bereich entsprechend der geschätzten Phasendrehung und dem Abstand von der Trainingssequenz durch Demodulation bestimmt wird und über die mittlere

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Phasenänderung die Änderung der Kanalimpulsantwort ermittelt wird.

4. Verfahren nach einem der vorherigen Ansprüche, dadurch gekennzeichnet, daß mit der geschätzten Phasendrehung durch entsprechende Mittelung der Versatz der Trägerfrequenz bestimmt wird und über einen Regelkreis die Trägerfrequenz reguliert wird.

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Fig. 1

Fig. 2

Fig. 3

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(54) **Zeitmultiplex-Verfahren zur Bestimmung der mittleren Phasenänderung eines Empfangssignals.**

(57) Die Erfindung betrifft ein Zeitmultiplex-Verfahren, mit dem die mittlere Phasenänderung eines Empfangssignals bestimmt wird. Im Mobilfunk ist die Kanalimpulsantwort durch den Dopplereffekt zeitlich veränderlich. Es stellt sich das Problem, diese Änderungen in einem Empfänger zu schätzen und zu berücksichtigen. Es wird ein einfaches Verfahren angegeben, welches die Änderungen der Kanalimpulsantwort durch eine mittlere Phasenänderung beschreibt. Das neue Verfahren ist für TDMA Systeme wie zum Beispiel das GSM System geeignet.

Fig. 2

EP 0 534 399 A3

EINSCHLÄGIGE DOKUMENTE

Kategorie	Kennzeichnung des Dokuments mit Angabe, soweit erforderlich, der maßgeblichen Teile	Betrifft Anspruch	KLASSIFIKATION DER ANMELDUNG (Int. Cl.5)
A	FREQUENZ Bd. 44, Nr. 7/8, Juli 1990, BERLIN DE Seiten 217 - 221 PLAGGE ET AL. 'Neues Verfahren zur Messung der Kanalstossantwort und Trägersynchronistion in digitalen Mobilfunktionen' * Zusammenfassung * * Seite 220, rechte Spalte - Seite 221, rechte Spalte, Zeile 5 * ---	1-4	H04B7/01
A	"Performance of Viterbi Equalisers for the GSM System". Second IEE National Conference on Telecommunications. 2-5-April 1989. Lopes p 61-66 * Seite 61, linke Spalte, Zeile 13 - Zeile 38 * * Seite 61, rechte Spalte, Zeile 20 - Seite 62, linke Spalte, Zeile 21 * ---	1-4	
A	"Design of a digital receiver for the GSM cellular system." Treizieme Colloque sur le Traitement du Signal et des Images. Juan-Les Pins. 16-20 September 1991. & Benelli et al. p477-480. Absatz 4. "A Digital Receiver for the GSM System" * Seite 479 * -----	1-4	RECHERCHIERTE SACHGEBIETE (Int. Cl.5) H04B H04L

Der vorliegende Recherchenbericht wurde für alle Patentansprüche erstellt

Recherchenort	Abschlußdatum der Recherche	Prüfer
DEN HAAG	22 APRIL 1993	GOULDING C.A.

(54) **Compensation of doppler errors, particular in cellular networks**

(57) In accordance with one embodiment of the invention, a method of compensating for Doppler error in a wireless communications system employing Viterbi decoding comprises the steps of: for each signal sample in a first predetermined-sized grouping of received signal samples, performing a parallel Viterbi update and short symbol decode; and for a second predetermined-sized grouping, forming by pipeline processing an estimate of the Doppler error in accordance with the parallel short traceback decoding performed for the first grouping, and adjusting each signal sample in the second grouping in accordance with the estimated Doppler error.

In accordance with another embodiment of the invention, a Viterbi traceback reconstructed signal sample index comprises: a state counter, a traceback shift register (TBSR); a signal reconstruction table; and a comparator coupled in a configuration so as to provide the sign bit to the traceback shift register from a comparison of binary digital signals. The state counter is coupled so as to provide digital signals to the traceback shift register and the traceback shift register is coupled so as to provide digital signals to the signal reconstruction table.

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Description

Technical Field

The present invention relates to communications and, more particularly, to wireless communications systems.

Background of the Invention

Several standards are being employed with respect to signaling standards for digital cellular telephony worldwide. One such standard is Europe's global system for mobile communications (GSM), described, for example, in the ETSI/GSM Specifications. Likewise, another standard, employed in the United States, is IS54. These signaling standards are typically intricate. Typically, a signal burst is transmitted via a wireless medium comprising a plurality of differentially encoded symbols, such as bits or binary digital signals. The encoded symbols are typically transmitted using phase shift keying in the baseband, such as, for example, minimum phase shift keying (MSK). This baseband signal form of the signal burst is typically then transmitted via a wireless medium using a radio frequency (RF) carrier.

In wireless communications, such as digital cellular telephony, transmission and reception typically occurs between stations, at least one of which is in motion. Furthermore, the velocity of the mobile station may change over time. As is well-known, such relative motion between transmitting and receiving stations, such as between a mobile station and a base station, may result in a Doppler shift of the frequency of the signal being transmitted. This Doppler shift may therefore result in a phase or frequency error in the received signal. Because of such a Doppler shift, the integrity of the signal being transmitted may be corrupted at the receiving end of the communications system.

Although various processes for compensating for such errors due to a Doppler shift are known, one problem associated with conventional approaches is the impact upon available signal processing resources at the receiving station. Typically, a receiving station, such as a mobile station, has limited signal processing capability. Therefore, exhaustive approaches to performing Doppler calculations and signal correction may exceed or at least bottleneck available resources when such resources are needed to continually process additional signals. A need therefore exists for a method of compensating for Doppler error while reducing associated bottlenecks for the available signal processing resources of a receiving station.

Summary of the Invention

In accordance with one embodiment of the invention, a method of compensating for Doppler error in a wireless communications system employing Viterbi decoding comprises the steps of: for each signal sample in a first predetermined-sized grouping of received signal samples, performing a parallel Viterbi update and short symbol decode; and for a second predetermined-sized grouping, forming by pipeline processing an estimate of the Doppler error in accordance with the parallel short traceback decoding performed for the first grouping and adjusting each signal sample in the second grouping in accordance with the estimated Doppler error.

In accordance with another embodiment of the invention, a Viterbi traceback reconstructed signal sample index comprises: a state counter, a traceback shift register (TBSR); a signal reconstruction table; and a comparator coupled in a configuration so as to provide the sign bit from a comparison of binary digital signals to the TBSR. The state counter is coupled so as to provide digital signals to the TBSR and the TBSR is coupled so as to provide digital signals to the signal reconstruction table.

Brief Description of the Drawings

The invention will be described by way of example with reference to the accompanying drawings in which:

FIG. 1 is a schematic diagram of one embodiment of a system employing a method of compensating for Doppler error in a wireless communications system in accordance with the invention.

FIG. 2 is plot of complex signals in the Inphase-Quadrature (I-Q) plane illustrating a phase offset that may be attributable to, for example, Doppler error.

FIG. 3 is a schematic diagram illustration of one embodiment of a signal burst or transmission burst, such as for GSM.

FIG. 4 is a flowchart illustrating one embodiment of a method of the invention.

FIG. 5 is a table illustrating calculations that may be performed by an embodiment of a method of the invention.

FIG. 6 is a block diagram illustrating an embodiment of a Viterbi traceback reconstructed signal sample index in accordance with the invention that may, for example, be employed by an embodiment of the invention.

FIG. 7 is a table illustrating automatic frequency correction (AFC) calculations that may be performed by an embodiment of a method of the invention.

FIG. 8 is a table illustrating a sequence of Viterbi operations in parallel with pipelined digital signal processor (DSP) operations that may be performed by an embodiment of a method of the invention.

Detailed Description

FIG. 3 illustrates a signal burst or transmission burst, such as may be employed in a wireless communications system, although the invention is not limited in scope to a signal burst having this particular format or

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structure. As illustrated in FIG. 3, the signal burst or transmission burst illustrated comprises a predetermined number of binary digital signals or bits. In this particular embodiment, each frame includes, in succession, a series of successive predetermined starting bits, a predetermined number of successive binary digital signals to be transmitted, a series of successive predetermined training bits, a second predetermined number of successive binary digital signals to be transmitted, and a series of successive predetermined ending bits. For GSM, for example, there are three starting and three ending bits, 58 bits in both portions of the signal burst comprising binary digital signals to be transmitted, and 26 training bits, referred to in this context as the "midamble," for a total of 148 bits per frame or signal burst. Of course, the invention is not restricted in scope to GSM.

As is well-known, GSM uses a form of signal modulation in the baseband known as Gaussian Minimum Phase Shift Keying (GMSK). GMSK is described in more detail in Digital Phase Modulation, by J. B. Anderson, T. Aulin and C. E. Sundburg, 1986, available from Plenum, although, of course, the invention is not restricted in scope to GMSK or even to MSK. For example, in IS54, an alternate baseband modulation scheme is employed. In such baseband modulation schemes, such as MSK or GMSK, the bit or binary digital signal stream to be transmitted, such as a signal burst, is differentially encoded, e.g., baseband modulated to produce a positive or negative phase shift representing one or more binary digital signals in the signal burst being transmitted. As previously described, this phase shift modulated baseband signal may then be applied to a radio frequency (RF) carrier for transmission via a wireless medium. Therefore, at the receiving end of the communications system, after downconversion and signal sampling, the binary digital signals in the signal burst being transmitted may be obtained by (1) a process, referred to as "derotation" in this context, applied to each signal sample in the signal burst and (2) then passing the derotated signal sample through a minimum least squares error (MLSE) equalizer. In this context, the term "differential encoded digital symbol" refers to a complex signal or signal sample at the receiving end of the communications system. The binary digital signals to be sent are transmitted as an analog signal via the modulation scheme employed in the baseband. The analog signal is then sampled at the receiving end in the baseband to provide the complex signal or signal sample. Depending on the modulation scheme employed, a symbol to be transmitted may comprise a predetermined set of one or more binary digital signals or bits. Furthermore, regarding "derotation," for GSM, for example, a rotation of 90° may be applied to each differentially encoded symbol or signal sample in the signal burst transmitted via the wireless medium, such as by signal multiplication in the baseband of each differentially encoded symbol or signal sample by

$$e^{\pm j\frac{\pi}{2}k},$$

where k=0, 1, 2, 3........ Of course, the invention is not restricted in scope to a signaling scheme employing a particular direction of rotation or derotation. The direction will depend, at least in part, on the particular signal modulation scheme employed. Likewise, the phase shift applied to "derotate" the baseband signal will depend on the modulation scheme employed. Nonetheless, as previously indicated, the Doppler effect due to the relative motion between the transmitting and receiving station, such as between a mobile station and a base station, may result in phase rotation error in the received signal in comparison with the signal transmitted. This is illustrated schematically in the Inphase-Quadrature (I-Q) plane in FIG. 2.

As illustrated, a Doppler shift may result in a frequency error in the received signal that may translate into a phase offset error in the complex signal obtained at the receiving end of the communications system. This offset error may also appear in the derotated signal obtained as well. In FIG. 2, the phase offset error is denoted by the phase difference, dθ, between Z and Ẑ. Z denotes the actual signal transmitted including a Doppler phase shift error and Ẑ denotes the Doppler corrected received signal. Likewise, Z' and Ẑ' denote these respective signals after derotation in the I-Q plane.

Various approaches to compensating the received signals for this phase offset error are known, such as described in "Two Stage Doppler Phase Corrected TCM/DMPSK for Shadowed Mobile Satellite Channels," by P. J. Mehane, appearing in IEEE Trans. on Communications, vol. 41, No. 8, August 1993, herein incorporated by reference. However, as previously indicated, such approaches are time-consuming and may also "bottleneck" significant signal processing capability in an environment having limited resources. A method of compensating for Doppler error in a wireless communications system in accordance with the invention to reduce such bottlenecks involves parallel and pipelined signal processing utilizing the computational resources available, such as, for example, a digital signal processor (DSP). For a method of compensating for Doppler error in a wireless communications system in accordance with the invention, portions of the signal processing to be performed may be segmented and "offloaded" to another processor or coprocessor. These segmented portions of the signal processing may be performed in advance by the coprocessor in a parallel fashion while the digital signal processor is also performing pipelined signal processing substantially in tandem. In this context, the term "pipelining" or "pipelined signal processing" refers to signal processing performed in a predetermined number of separate segments or stages. In such "pipelined signal processing," the processing result of a particular stage or segment is employed in the processing performed by the next stage or segment after the

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particular stage. The details of this approach will become clear in the discussion that follows.

This approach may be illustrated at least in part by the block diagram in FIG. 1. In this particular embodiment, a digital signal processor (DSP) 170 has embedded within it a Viterbi decoder 110. An example of such a digital signal processor is the DSP1618 available from AT&T Corp., which includes an embedded error correction coprocessor (ECCP) operating as Viterbi decoder 110, described in the preliminary data sheet, dated February 1994, available from AT&T Corp., herein incorporated by reference, although the scope of the invention is not limited in this respect. As illustrated, DSP 170 obtains the received signal burst. It will, of course, be appreciated that some preprocessing has typically been performed on the signal burst, such as downconversion, analog-to-digital conversion, and "derotation." Thus, each symbol in the received signal burst takes the form of a complex digital signal in the Inphase-Quadrature (I-Q) plane that itself represents one or more binary digital signals being transmitted, as previously described.

As illustrated, and as is well-known, the signal burst may be applied to a Viterbi decoder, such as Viterbi decoder 110, one signal sample at a time in order to obtain the transmitted signal burst based on the received signal burst. Viterbi decoding is well-known and described in, for example, Digital Communications, by E. Lee and D. Messerschmitt, available from Kluwer Academic Publishers, 1992, Digital Communications by Satellite, by Bhargava, Haccoun, Matyas, and Nuspl, available from John Wiley & Sons, Inc., 1981, and Digital Communications by Satellite, by J. J. Spilker, Jr., available from Prentice-Hall, Inc., 1977, all of which are herein incorporated by reference. Viterbi decoding is likewise described in "Maximum Likelihood Sequence Detection in the Presence of Intersymbol Interference," by G. D. Forney, Jr., and available in IEEE Trans. on Information Theory, IT-18(3): 363-378, May, 1972, and "The Viterbi Algorithm," IEEE Proceedings, March, 1973, 268-278, herein incorporated by reference. As illustrated in FIG. 1, however, prior to being applied to Viterbi decoder 110, each signal sample is adjusted by a phase offset provided by Doppler phase 150. Likewise, as illustrated in FIG. 1, this phase offset is obtained by Doppler phase 150 based on a prior predetermined-sized grouping of signal samples of the received signal burst that has been delayed by time delay 160 and compared with a signal estimate of the symbols transmitted for that prior grouping provided by Viterbi traceback reconstructed signal sample index 140. Delay 160 is introduced to ensure that the appropriate grouping of received signal samples is compared with the appropriate signal sample estimates. Channel estimate 120 provides signals to signal sample index 140 so that signal reconstruction may be performed. Signal reconstruction may be performed in accordance with any method known. Signal sample index 140 obtains signals from Viterbi decoder 110 based on a process performed by the decoder des-

ignated in FIG. 1 as a "short traceback" or "short decode." In this particular embodiment, a short traceback refers to a traceback of length one, although short tracebacks of length greater than one may also be employed, such as described, for example, in aforementioned patent application serial no. 08/152531, entitled "Variable Length Tracebacks." Likewise, as illustrated in FIG. 1, Viterbi decoder also provides signals based on a process referred to as a "long traceback" or "long decode." Thus, Viterbi decoder 110 has the capability to decode the received signal burst in order to determine the binary digital signals that have been transmitted. However, a Viterbi decoder for this embodiment of a method of compensating for Doppler error in a wireless communications system has the ability to perform at least two processes, one referred to as a "long traceback" and another referred to as a "short traceback."

One aspect of a method of compensating for Doppler error in a wireless communications system is a parallel and pipelined processing approach between Viterbi decoder 110 and DSP 170 to reduce bottlenecks that DSP 170 might typically encounter during signal processing. More specifically, and as described in more detail hereinafter, in processing a signal burst, such as illustrated in FIG. 3, for example, for a predetermined-sized grouping of the received differentially encoded symbols, referred to in this context as an intermediate grouping, in this particular embodiment each signal sample is provided to Viterbi decoder 110 in order to perform a Viterbi update and short traceback. Furthermore, in parallel with that process, for the immediately succeeding grouping of complex signal samples in the signal burst, referred to in this context as the second grouping, digital signal processor 170 forms by pipelined processing an estimate of the Doppler error in accordance with the short traceback previously performed for each signal sample in the grouping preceding the grouping currently being applied to the Viterbi decoder, referred to in this context as the first grouping. Likewise, the digital signal processor will adjust or correct each complex signal sample in the succeeding or second grouping in accordance with the estimated Doppler error. This is illustrated in FIG. 1, for example, in which a phase offset is applied to the received signal corresponding to the complex signals or signal samples previously described before the signal samples are applied to the Viterbi decoder. Thus, for this particular embodiment, at a given time, an intermediate grouping may be processed by the Viterbi decoder, a first grouping just processed by the Viterbi decoder may be processed by the DSP to estimate Doppler error and a second grouping about to be applied to the Viterbi decoder may be adjusted based on the Doppler error estimated from processing the first grouping. In this fashion, in this particular embodiment, the digital signal processor typically does not experience a bottleneck because processing by the digital signal processor continually processes in a pipelined fashion complex signal samples based on

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complex signal samples previously processed in parallel by the Viterbi decoder, while the Viterbi decoder in parallel continually processes additional complex signal samples that have been previously adjusted in phase by the DSP.

In a method of compensating for Doppler error in a wireless communications system in accordance with the invention, processing is applied to the received signal burst in advance of the pipelining previously described. For example, as previously indicated, the received burst signal is downconverted to provide a baseband signal and the baseband signal typically is converted from an analog signal to a complex signal in a quantized binary form, although the scope of the invention is not limited in this respect. This quantized binary digital signal represents a complex signal in the inphase-quadrature (I-Q) plane corresponding to one symbol of a plurality of differentially encoded symbols for a signal burst. Thus, the received signal burst comprises a plurality of complex signals or signal samples transmitted via a wireless communications system and may be provided to a processor, such as digital signal processor 170, in the signal form just described. As previously indicated, the transmitted signal burst has a substantially predetermined structure in which a subset of the binary digital signals being transmitted are known at the receiving end of the wireless communications system. As is well-known, this signal information may be employed to obtain an estimate of the communications channel and this channel estimate may then be used in further signal processing. In order to obtain this channel estimate, the received complex signal samples for the encoded symbols in the signal burst may first be "derotated" and, likewise, an automatic frequency correction (AFC), may be applied to compensate, for example, for frequency offset error attributable to the oscillator employed to downconvert the transmitted signal to a baseband signal, although the scope of the invention is not limited in this respect.

Channel estimation is well-known, such as described in "Design and Performance of Synchronization Techniques and Viterbi Adaptive Equalizers for Narrowband TDMA Mobile Radio," by G. D'Aria and V. Zingarelli, published in Nordic Seminar on Digital Land Mobile Radio Communication, 3rd Proceeding, Sept. 12-15th, 1988, Copenhagen, herein incorporated by reference, and a variety of signal processing techniques, such as digital signal processing, may be employed. Typically, the training bits in the signal burst being transmitted, may be employed to obtain an estimate of the communications channel. Likewise, "channel windowing" may be employed to determine the "maximum energy" portion of the signal and normalization and signal scaling may, likewise, be employed, although the invention is not limited in scope in this respect. Once channel estimation has been performed, as illustrated in FIG. 1, this channel estimate may be employed to perform signal reconstruction. Any known technique for signal reconstruction may be employed. In signal reconstruction, the channel estimate is employed to obtain an estimate of the complex signal sample transmitted for each encoded symbol capable of being transmitted. This may be obtained, for example, from the dot product of the channel estimate with vectors corresponding to the Viterbi states for the particular modulation scheme employed.

In this particular embodiment of a method of compensating for Doppler error in a wireless communications system in accordance with the invention, signal reconstruction is based at least in part on the assumption that the complex signal samples obtained at the receiving end of the wireless communications system correspond to locally maximum likelihood symbols also determined at the receiving end of the communications system. Therefore, reconstructed signals may then be employed in signal sample index 140, in conjunction with signals obtained from a short traceback performed by Viterbi decoder 110, to obtain an estimated signal provided to Doppler phase 150 for the transmitted symbol. For a predetermined-sized grouping of signal samples in the signal burst, this estimated signal or signal sample for each symbol may be compared with the received signal sample for the symbol in order to obtain an estimate of the phase offset, as described in more detail hereinafter.

As illustrated in FIG. 3, for this particular embodiment the signal burst may be divided or segregated into starting bits, ending bits, training bits and the binary digital signals to be transmitted. As illustrated in FIG. 1, the received signal samples corresponding to symbols are adjusted prior to being applied to Viterbi decoder 110. Therefore, initialization of the signal adjustments to be applied is desirable. In this particular embodiment of a method of compensating for Doppler error in a wireless communications system in accordance with the invention, it is desirable to use the training bits in order to initialize the Doppler portion of the processing for a variety of reasons. For example, the binary digital signals transmitted are known for the training bit portion of the signal burst. Thus, initialization using this portion of the signal burst provides a relatively high degree of reliability in terms of estimating Doppler error.

Although the scope of the invention is not limited in this respect, typically in wireless communications systems employing Viterbi decoding, the signal burst processing is divided into a forward portion and a backward portion, referred to in this context as "forward equalization" and "backward equalization." For this particular embodiment, signal processing is first applied to the "forward" portion, ranging from bit positions 77 to 147 in the signal burst illustrated in FIG. 3, for example. In FIG. 3, as illustrated, the initial bit position is numbered 0 and the remaining bit positions increase consecutively. One skilled in the art will nonetheless appreciate extension of the approach to the "backward" portion of the signal burst, such as to bits positions 65 to 0.

FIG. 4 is a flowchart depicting the pipelined and parallel processing approach employed in this particular

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embodiment of a method of compensating for Doppler error in a wireless communications system in accordance with the invention. As illustrated in FIG. 4, it is assumed that "forward" equalization is performed first. Nonetheless, the invention is not restricted in scope in this respect and, alternatively, backward equalization may be performed first. Next, as illustrated in FIG. 4 and described below, AI_{ACC} and AQ_{ACC} are initialized using the training bits. As previously Indicated, this is a useful aspect of a method of compensating for Doppler error in accordance with the invention in that initialization may be performed using transmitted binary digital signals that are known at the receiving end of the communications system.

In this particular embodiment, using the channel estimate previously obtained, the binary digital signals for positions 77-86 in the received signals burst are employed to obtain estimated signals or signal samples for the transmitted symbols and these estimated signal samples are compared with the signal samples actually received for these symbols to obtain the difference between respective inphase components and quadrature components of these compared signal samples. It is assumed in this context that the difference is largely attributable to Doppler error. Thus, the channel estimate is applied to known bit positions 77-82, 78-83, 79-84, 80-85, and 81-86 in the training bit portion of the signal burst. This provides 5 signal sample estimates corresponding to 5 successively transmitted symbols. These 5 signal sample estimates may now be compared with the complex signal samples received. The differences between the inphase components of the estimated and received complex signals are accumulated to obtain a ΔI_{ACC} estimate for this particular grouping of 5 signal samples. Likewise, a ΔQ_{ACC} estimate for this particular grouping is also obtained by accumulating the differences between the quadrature components, as illustrated in FIG. 4. This initialization approach is summarized by the first block in the flow chart shown in FIG. 4. In that block, Z_i represents the received signal samples with Doppler phase offset, whereas Sh_i^* represents the complex conjugate of the reconstructed signal sample based upon applying the channel estimate to the training bits, as just described. It will be appreciated that although in this particular embodiment a predetermined-sized grouping comprises 5 successive complex signal samples corresponding to successively transmitted symbols, the invention is not restricted in scope in this respect.

After bit position 86, the initialization portion has been completed and as illustrated in FIG. 4, a signal processing loop is employed in which Viterbi decoder 110 and DSP 170 in this particular embodiment perform signal processing in parallel so that a pipelined approach maybe employed for the signal processing performed by the DSP, as previously described. In particular in this embodiment, as illustrated in FIG. 4, DSP 170 employs ΔI_{ACC} and ΔQ_{Acc} obtained based on process-

ing of the immediately preceding grouping of complex signals to perform Doppler calculations and, likewise, using these calculations, performs Doppler signal corrections of the complex signals in the succeeding grouping in the signal burst i.e., bit positions 92-96 in this particular example for this particular embodiment, before those signals are provided to the Viterbi decoder. At the same time, the Viterbi decoder is performing Viterbi updating and short traceback decoding for the grouping beginning with bit position 87 for this particular embodiment. This Viterbi updating and short traceback decoding is performed a symbol or signal sample at a time and after processing upon each signal sample, the results obtained by the Viterbi decoder are employed by DSP 170 to compute ΔI_{ACC} and ΔQ_{ACC} for the grouping being processed by Viterbi decoder 110 in this particular embodiment. Likewise, in this particular embodiment, the Viterbi decoder performs a Viterbi update by performing a long traceback. The Viterbi decoder then performs a short traceback, of length one in this embodiment, as illustrated in FIG. 4. Likewise, once processing of this grouping by the Viterbi decoder is complete, then the DSP may be employed to perform the Doppler calculations and signal corrections for another grouping of signal samples based on the signal processing by the Viterbi decoder for this grouping while the Viterbi decoder is again performing Viterbi updates and short traceback decoding for the grouping of signals immediately after this grouping. Thus, it will now be appreciated by one skilled in the art that by employing this parallel processing structure, the Viterbi decoder continues to perform Viterbi updates and short traceback decodes while in parallel the DSP operates in a pipeline fashion to perform Doppler calculations and signal corrections for a grouping of signals based on the recently completed Viterbi processing of a prior grouping.

As illustrated in FIG. 4, eventually forward equalization is complete, in which case Doppler calculations and signal corrections are performed for the final grouping in the particular signal burst. Likewise, once this is complete, the process may be repeated for backward equalization. Ultimately, once the backward equalization is performed, the accumulated Doppler error from the backward equalization and the forward equalization may be combined to perform automatic frequency correction (AFC) for the next signal burst to be processed, as explained in more detail hereinafter. This automatic frequency correction update is performed because the Doppler signal corrections may be employed to assist in tracking phase offset not easily corrected by employing other automatic frequency correction techniques.

Various approaches to the Doppler calculations and signal corrections are possible and the invention is not restricted in scope to any particular approach. Likewise, the invention is not restricted in scope to a particular allocation of Doppler calculations or signal corrections to successive parallel Viteioi operations while processing signal samples in a particular grouping. Nonetheless, in

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this particular embodiment, 5 successive operations of Viterbi updating and short traceback decoding are employed in each grouping. On each parallel Viterbi operation, the DSP performs a different portion of the Doppler calculations and signal corrections. Likewise, in this embodiment, on each parallel operation, once the Viterbi update and short traceback decoding are performed, the DSP recomputes the accumulated ΔI and ΔQ, i.e., ΔI_{ACC} and ΔQ_{ACC}, for the grouping being processed by the Viterbi decoder based on the short traceback decoding just performed by the decoder. Signal processing by parallel Viterbi operations is summarized by the table provided in FIG. 5. For example, on the first Viterbi parallel operation, the DSP takes ΔI_{ACC} and ΔQ_{ACC} from the preceding grouping and obtains an estimate of the change in the phase offset from the inverse arctangent. Likewise, on the second parallel operation, the DSP obtains the average phase offset. This average is obtained from an accumulation of the phase offsets obtained for each grouping in the signal burst processed so far adjusted by the number of groupings processed. Likewise, on the third parallel operation, the accumulated phase offset is estimated based on the prior accumulated phase offset adjusted by a "weighting" function of the average differential phase offset. It will, of course, be appreciated that this weighting function may also be modified adaptively in alternative embodiments or omitted entirely. Likewise, on the fourth parallel operation, the sine and cosine of the estimated phase offset for this particular grouping is obtained. Finally, on the fifth parallel operation, Doppler signal corrections are made to the next grouping about to be processed by the Viterbi decoder, such as illustrated in FIG. 1. It will now be appreciated that this particular approach to pipelining the DSP with the Viterbi decoder processing in parallel introduces a slight lag in the Doppler signal corrections performed. It will be appreciated, however, that between successive groupings within a signal burst the impact of such a lag should not be significant. Furthermore, due to the successive nature of the parallel Viterbi operations, the effect of any slight lag should remain relatively fixed or stable over the signal burst.

After the Viterbi decoder has performed a Viterbi update and the short traceback for the next signal sample, such as indicated in FIG. 4, signals are provided to the DSP by the decoder so that ΔI and ΔQ may be obtained for the particular signal sample and then ΔI_{ACC} and ΔQ_{ACC} may be processed by the DSP. More particularly, once a short traceback is completed by the Viterbi decoder, an estimate of the complex signal transmitted is obtained using a Viterbi traceback reconstructed signal sample index in accordance with the invention, as described in more detail hereinafter, although the scope of a method of compensating for Doppler error in accordance with the invention is not limited in this respect. This estimate may then be provided to the DSP and the DSP may process the dot product of the complex conjugate of this signal sample estimate for the symbol with the

actual signal sample for the symbol in order to obtain ΔI and ΔQ, although the scope of the invention is not limited in this respect. Likewise, ΔI and ΔQ for each of the signal samples obtained, as just described, may be accumulated so that once the final signal sample in a particular grouping is processed, ΔI_{ACC} and ΔQ_{ACC} may be obtained for that grouping. As previously indicated, a similar approach to processing ΔI_{ACC} and ΔQ_{ACC} may be employed in conjunction with the training bits for the purposes of initialization.

FIG. 8 is a table illustrating the sequence of parallel Viterbi operations that may be performed by an embodiment of a method of compensating for Doppler error in a wireless communication system in accordance with the invention. The table in FIG. 8 illustrates the parallel Viterbi operations with the pipelined DSP processing for bit positions 87 to 92. As illustrated, this embodiment employs 5 parallel operations or operation cycles, although the invention is not limited in scope in this respect. Therefore, on the Viterbi operation performed by Viterbi decoder 110, in parallel the DSP performs a different portion of the Doppler calculations and signal corrections, as previously described and as illustrated in FIG. 8. An advantage of this pipelined parallel processing approach is that the utilization of the DSP and the Viterbi decoder is improved in comparison with alternate approaches. More particularly, less idle time occurs for the processor.

Although the invention is not restricted in scope to performing automatic frequency correction or to this particular technique of automatic frequency correction, in one embodiment, after a phase offset, $\theta(n)$, has been accumulated based upon forward equalization attributable to Doppler error, denoted in this context as $\theta_F(N)$ and, likewise, $\theta_B(N)$ has been accumulated based upon backward equalization, as illustrated in FIG. 4, these phase offsets may be employed to perform automatic frequency correction (AFC) for later signal bursts to be received. In this particular embodiment, as illustrated by the table shown in FIG. 7, the phase offset $[\theta_F(N) - \theta_B(N)]$ may be accumulated over a predetermined number of signal bursts and then employed to update the phase compensation employed to perform automatic frequency correction.

As indicated previously, Viterbi decoding is well-known and employed in a variety of technologies, such as in wireless communications. As is well-known, one aspect of Viterbi decoding relates to performing Viterbi updates or trellis decoding using received signal samples at the receiving end of a communications system. As received signal samples are processed by the Viterbi decoder, a "survival metric" is obtained for each Viterbi state in accordance with a Viterbi add-compare-select operation. After a predetermined number of signal samples have been processed, symbol decoding is performed in a manner referred to as a "long traceback" or "long decode" that reflects the most likely set of symbols to have been transmitted based upon the accumulated

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metrics.

In some situations, however, it may be desirable to perform a process referred to as a "short traceback." In this situation, it may be desirable to form a local estimate of the likely symbol to have been transmitted based upon recently available signal information. Although this estimate may not be as good as the estimate based upon a "long decode" or "long traceback;" nonetheless, in some circumstances, it may be necessary or desirable to have an early estimate available for signal processing purposes before a long traceback may be performed. One example of this desirability is in the context of performing Doppler error calculations and signal corrections. Typically, in these situations, it may be desirable or important to obtain the results of a short traceback in a timely and efficient manner. Typically, performing a process referred to as "traceback packing" as part of the traceback is a cumbersome process including relatively significant processing complexity. In this context, "traceback packing" refers to the process by which binary digital signals or bits indicated by the Viterbi decoding process to have been transmitted are concatenated for signal processing purposes.

FIG. 6 is a schematic diagram of embodiment 1000 of a Viterbi traceback reconstructed signal sample index in accordance with the invention providing a relatively efficient technique for performing traceback packing, particularly for a short traceback. As illustrated in FIG. 1, embodiment 1000 includes a counter or state counter 300, a traceback shift register (TBSR) 600, a signal reconstruction table 800, and a comparator 900 coupled in a configuration so as to provide the sign bit to the traceback shift register from a comparison of binary digital signals performed by comparator 900. As illustrated in FIG. 6, state counter 300 is coupled so as to provide digital signals to traceback shift register 600. Furthermore, traceback shift register 600 is coupled so as to provide digital signals to signal reconstruction table 800, via address decoder 700 for the embodiment illustrated in FIG. 6.

Viterbi traceback reconstructed signal sample index 1000 is intended for use in conjunction with a Viterbi decoder. As is well-known, typically in Viterbi decoding, upon processing of another received signal sample, a Viterbi update is performed in which stored signals, referred to in this context as "metrics," are updated. Once this Viterbi update has been performed, the stored metrics may be provided to comparator 900, as illustrated in FIG. 6, in order to perform a short traceback in accordance with the invention.

As illustrated in FIG. 6, MUXes 910 and 920 respectively couple registers 915 and 925 to the respective input ports of comparator 900. Although not explicitly illustrated in FIG. 6, these MUXes may be coupled, for example, to Viterbi decoder 110 or a RAM coupled to decoder 110 so that the "metrics" processed by Viterbi decoder 110 are available for comparison by comparator 900. In accordance with the Viterbi decoding proc-

ess, for a particular Viterbi state, two metrics are compared to determine the most extreme value metric for that particular Viterbi state. The extreme metric thereby obtained for that state may be stored for later comparison with the extreme metrics similarly obtained for the other Viterbi states. An embodiment of a Viterbi reconstructed signal samples index in accordance with the invention provides an efficient technique for performing this processing, as described in more detail hereinafter.

As illustrated in FIG. 6, state counter 300, in response to an externally-derived clock enable signal, may successively provide digital signals corresponding to the Viterbi states capable of being transmitted by the particular communications system. For example, in FIG. 6, four Viterbi states are illustrated using a two-bit counter, although the invention is not restricted in scope in this respect. As illustrated in FIG. 6, the counter provides these digital signals to traceback shift register 600. At substantially the same time, MUXes 910 and 920 provide two stored metrics to comparator 900 for the state indicated by counter 300, the stored metrics being derived from processing by Viterbi decoder 110. More particularly, in accordance with conventional Viterbi processing, these signals being compared represent the sum of a calculated branch metric and accumulated cost for the Viterbi state indicated by counter 300. Likewise, the sign bit obtained by comparator 900 based on the comparison of the two metrics is then stored in register 930. Depending on the extreme value of the metric obtained for this state, this sign bit may later be transferred to traceback shift register 600 as a "short traceback bit," as explained in more detail hereinafter.

As illustrated, the sign bit from this comparison also operates as enable signal EN1 to MUX 935. Signal EN1 operates, via MUX 935, to transfer the one extreme signal of the two signals respectively stored in registers 915 and 925 to register 940. Thus, register 940 operates as a "survival metric" register. Now, the contents of register 940, via MUX 920, may be compared with the contents of register 400, which stores the current value of the extreme metric for the Viterbi states processed up to this point. Comparator 900 then compares the "survival metric" value for the current Viterbi state as indicated by state counter 300, stored in register 940, with the current extreme metric value for the previously processed Viterbi states, stored in register 400. Assuming the "survival metric" for the current state is more extreme, i.e., greater or smaller depending upon the particular embodiment, the sign bit from comparator 900, denoted in FIG. 6 as EN2, enables the contents of register 940 to be loaded into register 400. Likewise, the contents of counter 300, indicating the current Viterbi state, is loaded as a digital signal into the portion of traceback shift register 600 denoted 610 in FIG. 6. Likewise, the sign bit stored in register 930 is loaded into the portion of traceback shift register 600 denoted as 620 and becomes the "short traceback bit." It will, of course, be appreciated that this process is repeated for each Viterbi state indicated by state

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counter 300. Thus, if a more extreme metric is obtained later, the contents of register 400 is loaded with that extreme metric signal value and binary digital signals representing the corresponding Viterbi state and short traceback bit are loaded in traceback shift register 600. It should now be clear to one of ordinary skill in the art that once counter 300 has provided signals to traceback shift register 600 corresponding to all the Viterbi states capable of being transmitted for the communications system, register 400 should contain the signal value for the most extreme metric of the Viterbi states and, likewise, traceback shift register 600 should contain binary digital signals representing the particular Viterbi state corresponding to that metric signal value and the associated short traceback bit. As illustrated in FIG. 6, the contents of traceback shift register 600 may now be provided to a signal reconstruction table 800 with memory location addresses that correspond to the concatenation of the Viterbi states in the form of a binary digital signal with a short traceback bit. The contents of the memory locations of signal reconstruction table 800 may contain the reconstructed signal sample values corresponding to the associated Viterbi states including a short traceback bit. Likewise, alternatively, as illustrated in FIG. 6, address decoder 700 may translate or decode the binary digital signals provided by traceback shift register 600 into the memory location address corresponding to the particular reconstructed signal sample value.

One advantage of a traceback shift register in accordance with the invention, such as the embodiment illustrated in FIG. 6, is that it provides a means to efficiently address a table of reconstructed signal sample values. By comparing "metrics," as previously described, the traceback shift register ultimately contains a binary digital signal that represents the binary digital signal most likely to have been transmitted via the communications system based at least in part upon recently available signal information. Thus, this binary digital signal may now be employed to address a memory storing estimated signal samples, i.e., reconstructed signal samples, for the binary digital signals capable of being transmitted by the particular communications system. In this particular embodiment, a 2-bit Viterbi state is concatenated with one short traceback bit assuming, for example, a three tap communications channel estimate is employed, although the invention is, of course, not restricted in scope in this respect.

It will now be appreciated that a Viterbi traceback reconstructed signal sample index in accordance with the invention offers advantages of speed, power savings and efficiency in comparison with traditional Viterbi traceback packing approaches. The desired reconstructed signal sample value may be quickly and efficiently addressed using a traceback shift register in accordance with the invention, as previously described. It will also now be appreciated that, although the embodiment shown in FIG. 6 includes a comparator in which

the extreme signal value of two binary digital signals, such as the smaller signal, is obtained, alternatively, this technique may be employed to find the larger of the two binary digital signals.

While only certain features of the invention have been illustrated and described herein, many modifications, substitutions, changes or equivalents will now occur to those skilled in the art. It is, therefore, to be understood that the appended claims are intended to cover all such modifications and changes as fall within the true spirit of the invention.

Claims

1. A method of compensating for Doppler error in a wireless communications system employing Viterbi decoding, said method characterized by the steps of:

     for each signal sample in a first predetermined-sized grouping of received signal samples,

     performing a parallel Viterbi update and short traceback decode; for a second predetermined-sized grouping,
     forming by pipeline processing an estimate of the Doppler error in accordance with the parallel short traceback decoding performed for the first grouping, and
     adjusting each signal sample in the second grouping in accordance with the estimated Doppler error.

2. The method of claim 1, where in the step of forming by pipeline processing an estimate of the Doppler error is characterized by forming the estimate by pipeline processing in parallel with performing a plurality of parallel Viterbi update and short traceback decodes for signal samples in an intermediate grouping between the first grouping and the second grouping.

3. The method of claim 2, wherein the second grouping is characterized by a grouping immediately after the intermediate grouping in the signal burst;
     the intermediate grouping comprising a grouping immediately after the first grouping in the signal burst.

4. The method of claim 2,

     wherein the second grouping is characterized by the grouping immediately preceding the intermediate grouping in the signal burst;
     the intermediate grouping comprising a grouping immediately preceding the first grouping in the signal burst.

5. The method of claim 2,
wherein the step of performing a parallel Viterbi update and short traceback decode is characterized by performing a short traceback decode with a Viterbi traceback reconstructed signal sample index.

6. A method of compensating for Doppler error in a wireless communications system characterized by the steps of:

receiving a signal burst transmitted via the wireless communications system, the transmitted signal burst comprising a first plurality of symbols including a second plurality of substantially predetermined symbols;
estimating the channel for the wireless communications system by processing the portion of the received signal burst corresponding to the second plurality of symbols;
forming reconstructed signals from the estimated channel;
for each signal sample in a first predetermined-sized grouping of received signal samples in the signal burst,

performing a parallel Viterbi update and short traceback decode in order to select a reconstructed signal sample, and
comparing the reconstructed signal sample with the corresponding derotated received signal sample; and

for a second predetermined-sized grouping,
forming by pipeline processing an estimate of the Doppler error in accordance with the signal sample comparisons for the first grouping, and
adjusting each signal sample in the second grouping in accordance with the estimated Doppler error.

7. The method of claim 6, wherein the step of forming by pipeline processing an estimate of the Doppler error is characterized by forming the estimate by pipeline processing in parallel with performing a plurality of parallel Viterbi update and short traceback decodes for signal samples in an intermediate grouping between the first grouping and the second grouping.

8. The method of claim 7, wherein the second grouping is characterized by a grouping immediately after the intermediate grouping in the signal burst;
the intermediate grouping comprising a grouping immediately after the first grouping in the signal burst.

9. The method of claim 7,
wherein the second grouping is characterized by a grouping immediately preceding the intermediate grouping in the signal burst;
the intermediate grouping comprising a grouping immediately preceding the first grouping in the signal burst.

10. The method of claim 6,
wherein the step of performing a parallel Viterbi update and short traceback decode is characterized by performing a short traceback decode with a Viterbi traceback reconstructed signal sample index.

11. The method of claim 6 further characterized by, prior to the steps of performing, comparing, forming, and adjusting for first and second predetermined-sized groupings, a step of initializing the estimate of the Doppler error is performed using the second plurality of substantially predetermined symbols.

12. A Viterbi traceback reconstructed signal sample index including a state counter, a traceback shift register, a signal reconstruction table, and a comparator, the signal sample index characterized by:

the comparator coupled in a configuration so as to provide the sign bit to said traceback shift register from a comparison of binary digital signals;
said state counter being coupled so as to provide digital signals to said traceback shift register; and
said traceback shift register being coupled so as to provide digital signals to said signal reconstruction table.

13. The Viterbi traceback reconstructed signal sample index of claim 12, characterized in that

said configuration includes a temporary traceback bit register coupled to said comparator so as to receive said sign bit; and
said temporary traceback bit register being coupled to said traceback shift register.

14. The Viterbi traceback reconstructed signal sample index of claim 13, characterized in that
said configuration is adapted to load the contents of said state counter and said temporary traceback bit register into said traceback shift register in response to an enabling signal provided by said comparator.

15. The Viterbi traceback reconstructed signal sample index of claim 12, in which said signal reconstruc-

a binary digital signal address decoder coupled to a memory;

said memory including binary digital signals *5* stored in memory locations addressed by said address decoder; and

said stored binary digital signals comprising reconstructed signal samples corresponding to predetermined states provided as binary digital *10* signals to said address decoder by said traceback shift register.

15

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55

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FIG. 1


```
                                                    170
                          ┌─────────────────────────────────┐
                          │                    110          │
                          │                     ┌──────────┐ │
  RECEIVED          ──────┼────────○────────────│ VITERBI  │─┼──── LONG DECODE
  SIGNAL                  │        │     -      │ DECODER  │ │
                          │  e^jθ(n+1)  PHASE   └──────────┘ │
                          │             OFFSET       │       │  SHORT
                          │                          │       │  DECODE
                          │         120─┐   ┌──────────┐     │
          160    ┌───────┐│             └───│ CHANNEL  │     │
           └─────│ DELAY │                  └──────────┘     │
                 └───────┘│                      │      140  │
                          │                      │  ┌─────────────────────┐
                          │                      │  │ VITERBI TRACEBACK   │
                          │                      └──│ RECONSTRUCTED SIGNAL│
                          │  ┌──────────┐           │ SAMPLE INDEX        │
                          │  │ DOPPLER  │           └─────────────────────┘
                          └──│ PHASE    │──────────────┘
                          │  └──────────┘                  │
                   RECEIVED       │      ESTIMATED          │
                   SIGNAL        150      SIGNAL            │
                          └──────────────────────────────────┘
```

100

FIG. 2

FIG. 3

```
|<-------------------- 148 BITS -------------------->|

+-------+----------------+---------------+----------------+-------+
|  000  |                |  "MIDAMBLE"   |                |  000  |
|   .   |<-- 58 BITS --->|<-- 26 BITS -->|<-- 58 BITS --->|   .   |
|   .   |                |               |                |   .   |
|   .   |                |               |                |   .   |
+-------+----------------+---------------+----------------+-------+
    0                                                        147
```
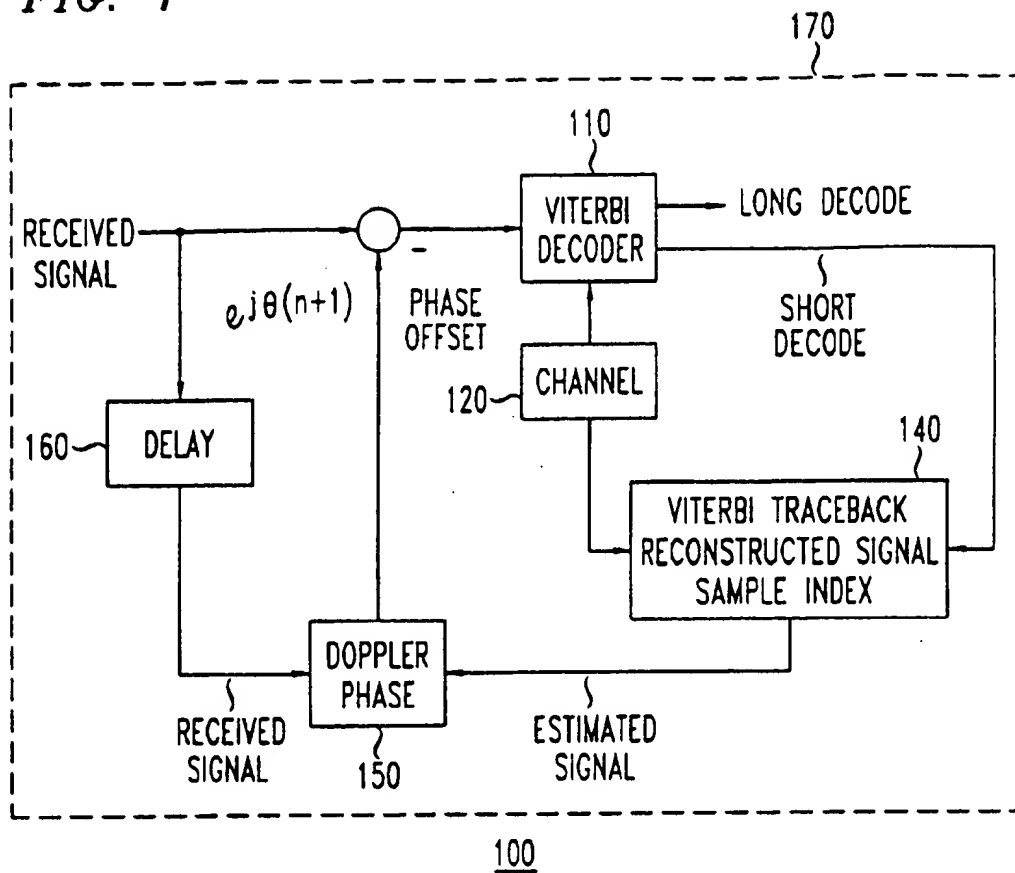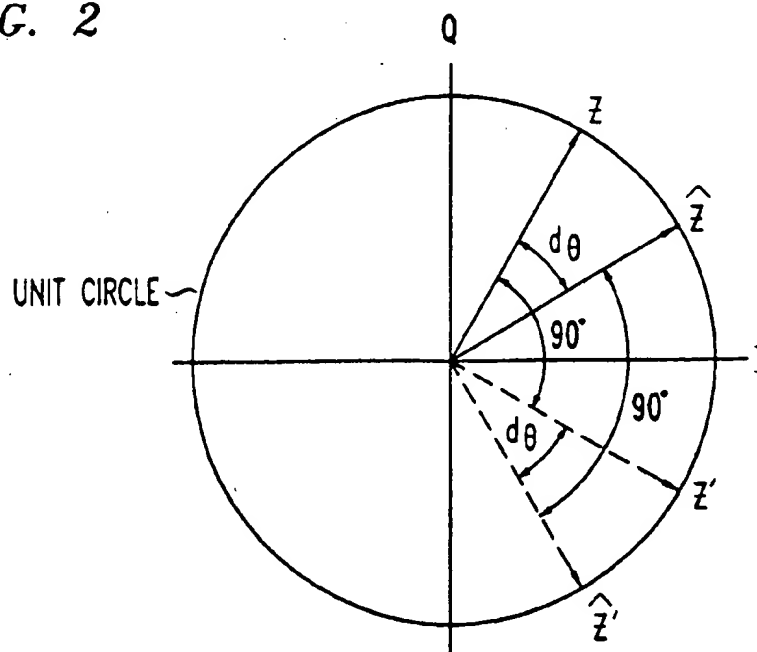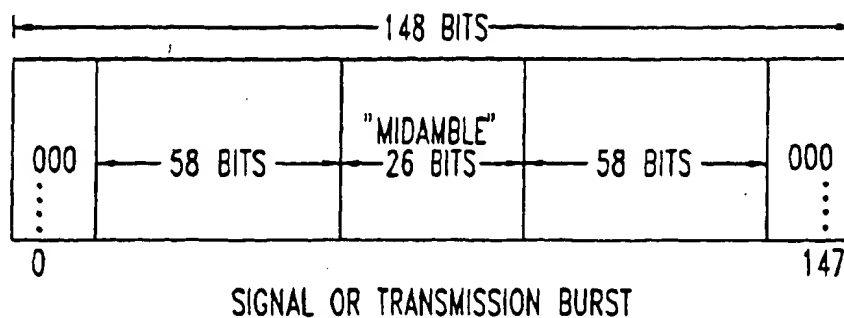
SIGNAL OR TRANSMISSION BURST

FIG. 5

INITIALIZATION

$$\Delta \text{ SUM} = 0$$
$$\text{COUNT} = 0$$
$$n = 0$$
$$\theta(o) = o$$

PARALLEL VITERBI OPERATION	PIPELINED DOPPLER CALCULATIONS AND SIGNAL CORRECTIONS
1	$\Delta\theta = \tan^{-1} \Delta Q/\Delta I$
2	$\Delta\text{ sum} = \Delta\text{ sum} + \Delta\theta$ $\text{count} = \text{count} + 1$ $\Delta_{AV} = \Delta\text{ sum}/\text{count}$
3	$\theta(n+1) = \theta(n) + \mu\,\Delta_{AV}$ $\mu = 0.3$
4	$\emptyset_1 = S_{in}\,\theta(n+1)$ $\emptyset_2 = C_{os}\,\theta(n+1)$
5	$I' = I\emptyset_2 + Q\emptyset_1$ $Q' = Q\emptyset_2 - I\emptyset_1$

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FIG. 4

ALL SYMBOLS DONE FOR FORWARD EQUALIZATION?

NO → B

YES ↓

SAVE $\theta_F(N)$

↓

PERFORM BACKWARD EQUALIZATION

↓

SAVE $\theta_B(N)$

↓

UPDATE AFC

DOPPLER INITIALIZATION

$$\Delta I_{acc} = \sum_{i=n}^{n+4} Re(Z_i \cdot Sh_i^*)$$

$$\Delta Q_{acc} = \sum_{i=n}^{n+4} Im(Z_i \cdot Sh_i^*)$$

DSP

BASED ON VITERBI UPDATES OF LAST GROUPING OF SYMBOLS, DSP PERFORMS DOPPLER CALCULATIONS AND SIGNAL CORRECTIONS IN PARALLEL WITH VITERBI DECODER OPERATIONS (SEE FIG.5 FOR PIPELINE SEQUENCE)

AFTER VITERBI PERFORMS ITS PROCESSING, THE DSP UPDATES $\Delta I_{acc}, \Delta Q_{acc}$

VITERBI DECODER

FOR NEXT SYMBOL VITERBI DECODER PERFORMS VITERBI UPDATE FOR LONG TRACEBACK FOLLOWED BY SHORT TRACEBACK
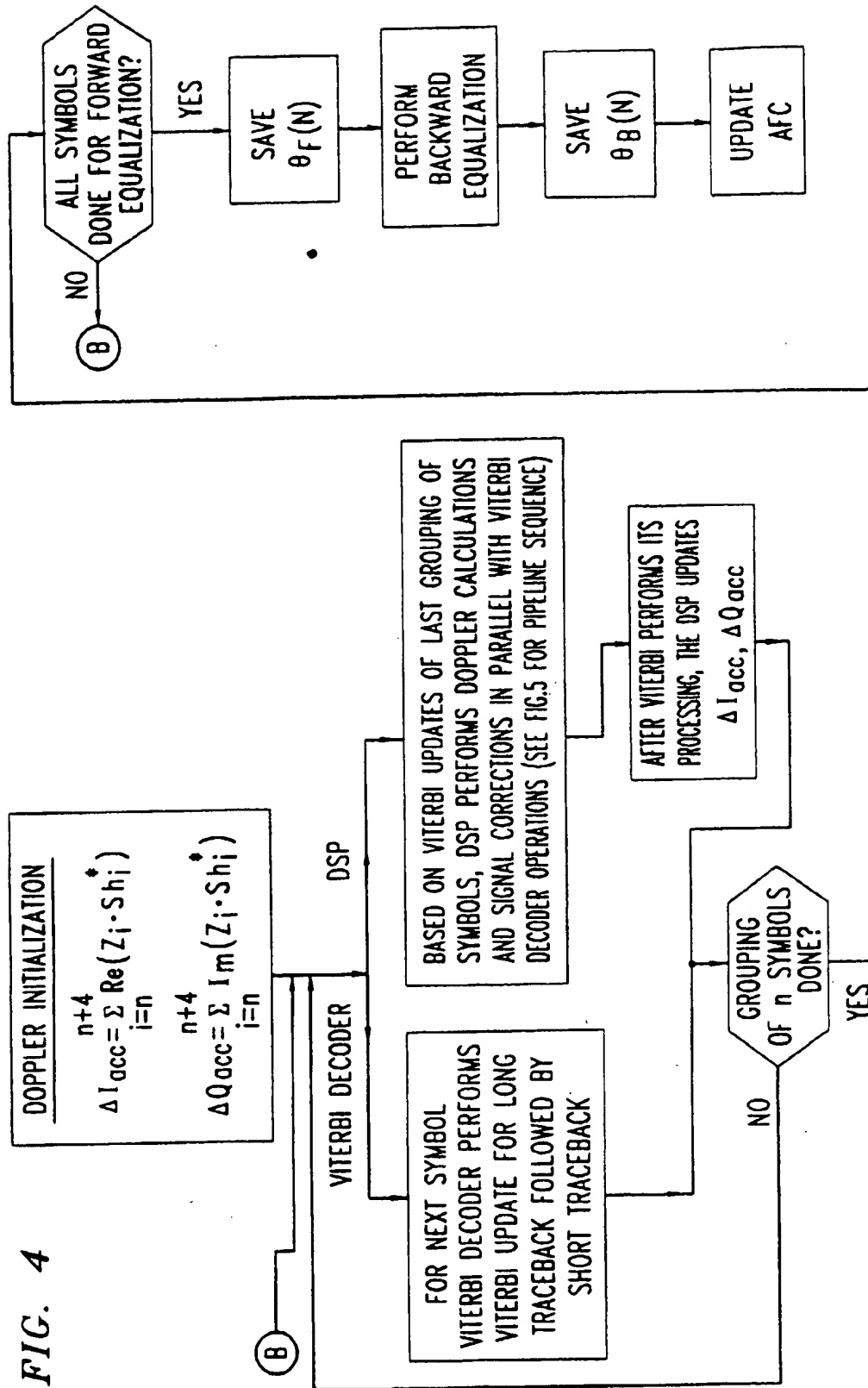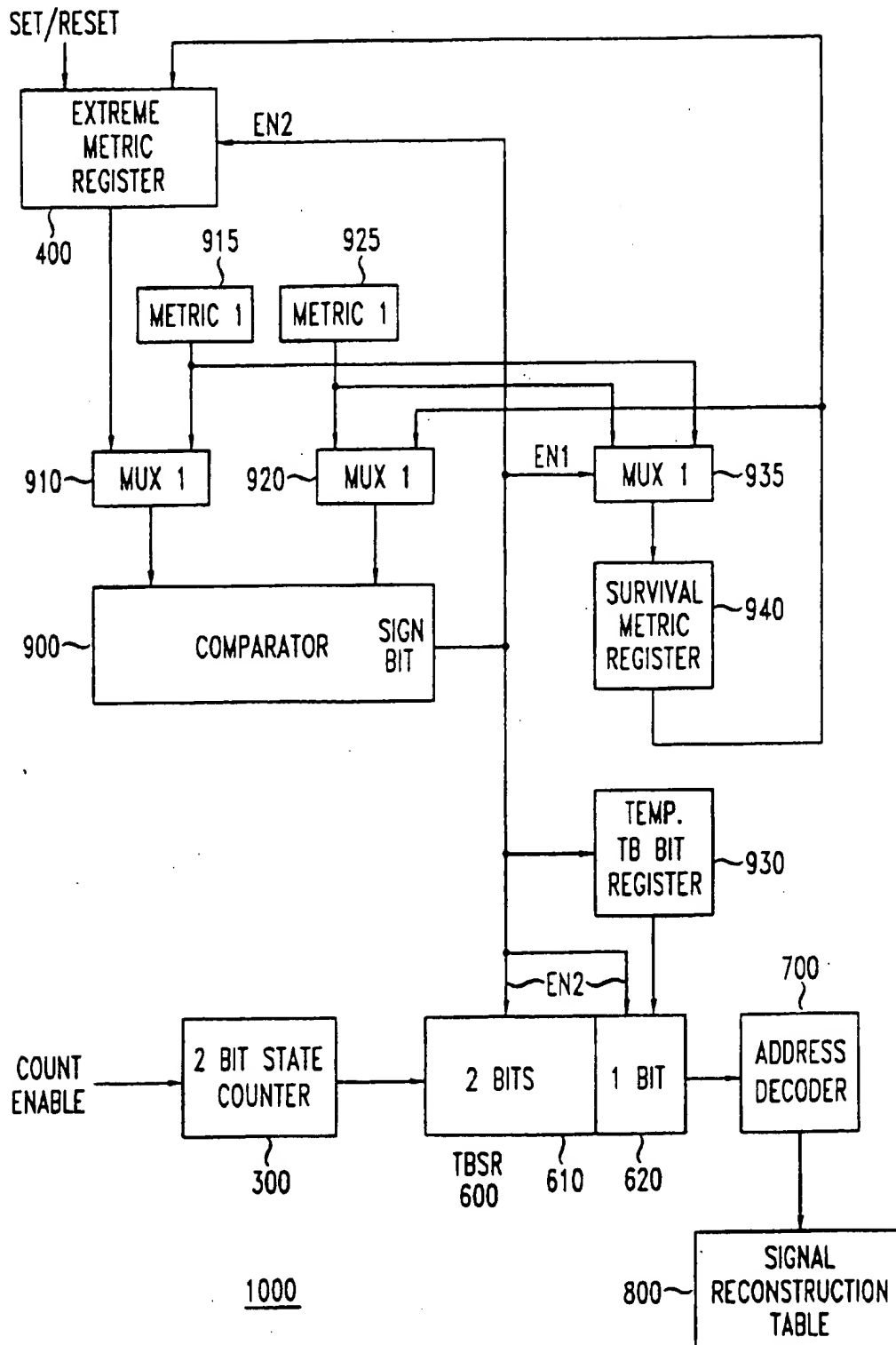
B

GROUPING OF n SYMBOLS DONE?

NO

YES

FIG. 6

FIG. 7

INITIALIZATION

BURST_COUNT = 0
PHASE_OFFSET = 0

AFTER COMPLETING A SIGNAL BURST

BURST_COUNT = BURST_COUNT + 1

$$\text{PHASE\_OFFSET} = \text{PHASE\_OFFSET} + \theta_F(N) - \theta_B(N)$$

IF (BURST_COUNT = K), THEN:

$$\left\{ \begin{array}{l} \text{AFC} = \text{AFC} + \text{PHASE\_OFFSET}/K \\ \text{BURST\_COUNT} = 0 \\ \text{PHASE\_OFFSET} = 0 \end{array} \right.$$

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FIG. 8

PARALLEL VITERBI OPERATIONS

	1	2	3	4	5
DSP: 87 / **VITERBI:**	Δθ — ΔIacc ΔQacc — XXX; UPDATE AND TRACEBACK; DSP: VITERBI:				
88		Δ SUM, COUNT, Δ AV — ΔIacc ΔQacc — XXX; UPDATE AND TRACEBACK; DSP: VITERBI:			
89			θ (n+1) — ΔIacc ΔQacc — XXX; UPDATE AND TRACEBACK; DSP: VITERBI:		
90				SIN θ (n+1), COS θ (n+1) — ΔIacc ΔQacc — XXX; UPDATE AND TRACEBACK; DSP: VITERBI:	
91					DOPPLER CORRECTION 92–96 — ΔIacc ΔQacc — XXX; UPDATE AND TRACEBACK
DSP: 92 / **VITERBI:**	Δθ — ΔIacc ΔQacc — XXX; UPDATE AND TRACEBACK				